

Microwaves & RF

THE HIGH SPEED ELECTRONICS GROUP

News

Previewing MTT-S
products and sessions

Design Feature

Make accurate
burst measurements

Product Technology

Smart clocks set
timing standards

Flexible Chip Set Arms 802.11 a/b/g WLANs



802.11

802.11

802.11

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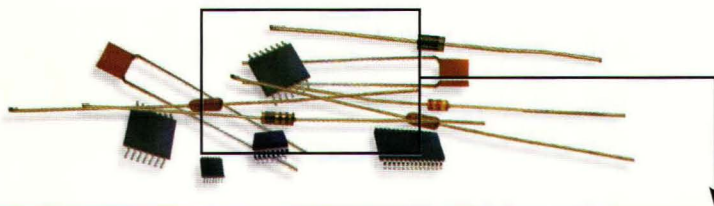


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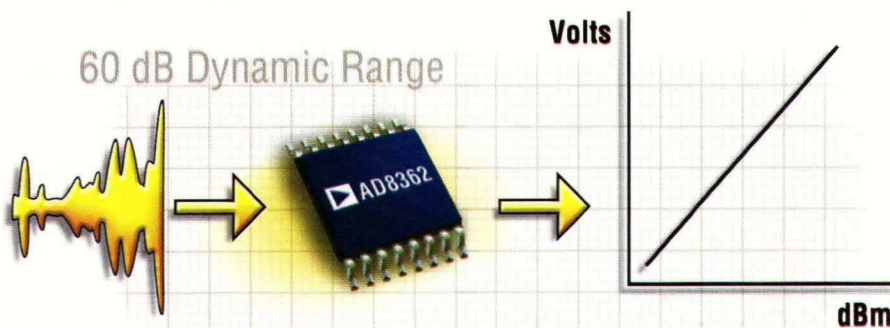
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**MTT-S Preview
Radar & Antennas
Issue**

Power measurement in multiple components.



RMS detection in a single IC.



AD8362 RMS Detector

The ability to measure power at RF frequencies is a tough challenge for next-generation wireless cellular equipment. The AD8362 is the only solution that computes RMS signal level with linear-in-dB output, and can measure signals with varying peak-to-peak average ratios up to 2.7 GHz and 60 dB dynamic range. This new RF IC from Analog Devices can provide RMS measurement of complex signals such as CDMA/W-CDMA, EDGE, and QAM simply and accurately in a single chip. **For more information on the AD8362 and the rest of our high-performance RF IC portfolio, call 1-800-ANALOGD or visit our website.**

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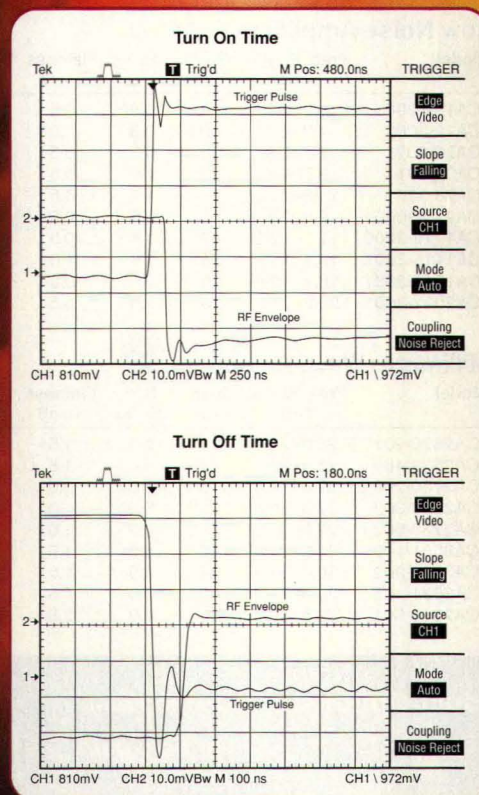
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Ultra Broadband Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA018-204	0.5-18.0	25	4.0	2.5	10	20	2.0:1	300
JCA218-506	2.0-18.0	35	5.0	2.5	15	25	2.0:1	400
JCA218-507	2.0-18.0	35	5.0	2.5	18	28	2.0:1	450
JCA218-407	2.0-18.0	30	5.0	2.5	21	31	2.0:1	500
JCA220-209	2.0-20.0	20	6.0	3.0	20	30	2.0:1	500

Power Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-P01	1.35-1.85	35	4.0	1.0	33	41	2.0:1	1000
JCA34-P02	3.1-3.5	40	4.5	1.0	37	45	2.0:1	2200
JCA56-P01	5.9-6.4	30	5.0	1.0	34	42	2.0:1	1200
JCA812-P03	8.0-12.0	40	5.0	1.5	33	40	2.0:1	1700
JCA1218-P02	12.0-18.0	22	4.0	2.0	25	35	2.0:1	700

Low Noise Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-1000	1.2-1.6	25	0.8	0.5	10	20	2.0:1	80
JCA12-3001	1.0-2.0	40	0.8	1.0	10	20	2.0:1	200
JCA23-302	2.2-2.3	30	0.8	0.5	10	20	2.0:1	80
JCA34-301	3.7-4.2	30	1.0	0.5	10	20	2.0:1	90
JCA78-300	7.25-7.75	27	1.2	0.5	13	23	2.0:1	120
JCA910-3000	9.0-9.5	25	1.3	0.5	13	23	1.5:1	150
JCA1112-3000	11.7-12.2	27	1.4	0.5	13	23	1.5:1	150
JCA1415-3001	14.4-15.4	35	1.6	1.0	14	24	2.0:1	200
JCA1819-3001	18.1-18.6	25	2.0	0.5	10	20	2.0:1	200
JCA2021-3001	20.2-21.2	25	2.5	0.5	10	20	2.0:1	200

Millimeter Wave Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA2629-201	26.0-29.0	19	5.0	1.5	5	15	2.0:1	100
JCA2629-401	26.0-29.0	35	5.0	1.5	5	15	2.0:1	200
JCA2730-205	27.5-30.0	15	5.0	1.0	15	25	2.0:1	200
JCA2730-302	27.5-30.0	26	5.0	1.0	8	18	2.0:1	150
JCA2730-502	27.5-30.0	43	5.0	1.0	8	18	2.0:1	200
JCA3031-102	30.0-31.0	18	5.0	1.5	8	18	2.0:1	100
JCA3031-302	30.0-31.0	34	5.0	1.5	8	18	2.0:1	200
JCA3031-405	30.0-31.0	40	5.0	1.5	15	25	2.0:1	400
JCA2640-301	26.5-40.0	30	5.0	2.5	0	10	2.0:1	160

Product Options:

- Limiting amp
- Variable gain control
- TTL switching
- Temperature compensation
- Alternate gain, N.F., power, VSWR levels
- Input/output isolators
- Waveguide interface

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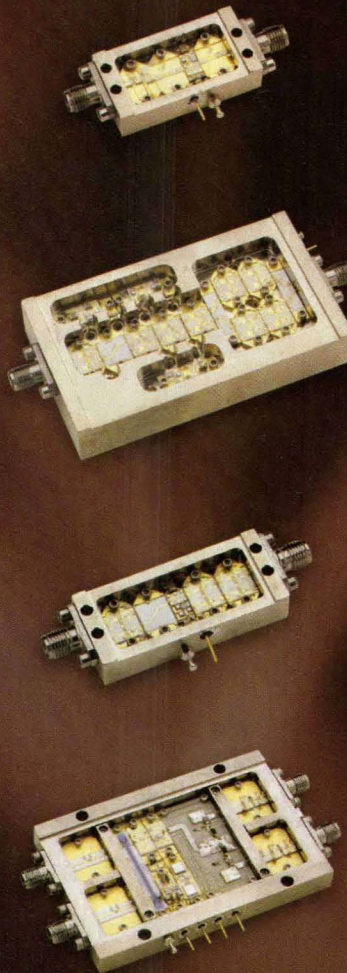
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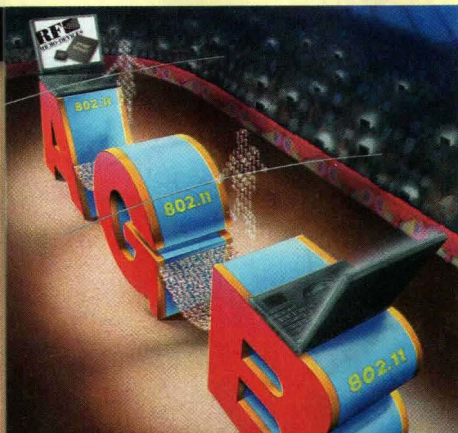
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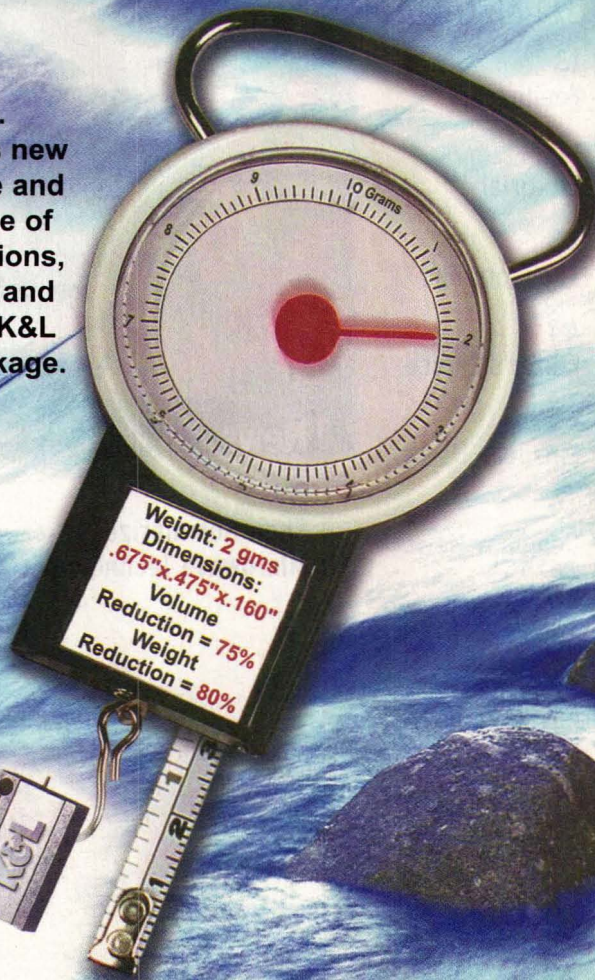


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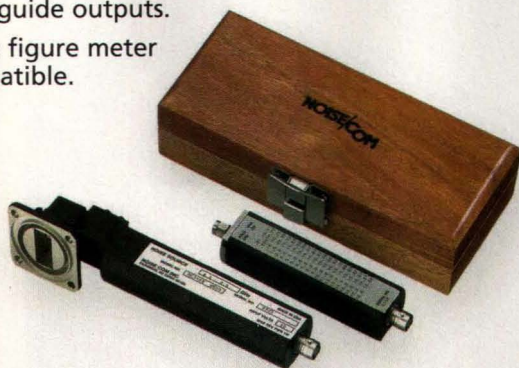
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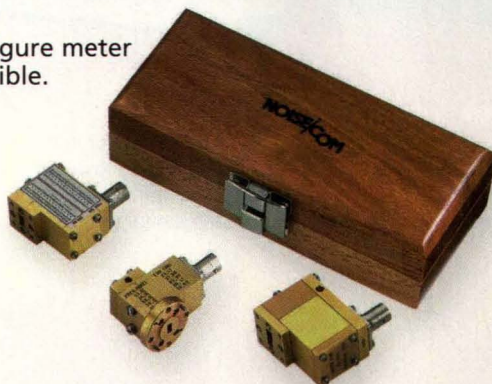
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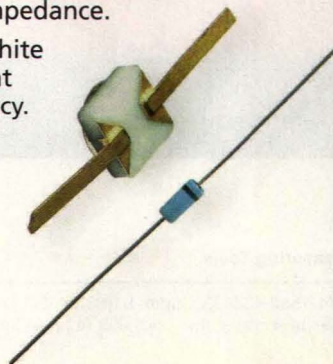
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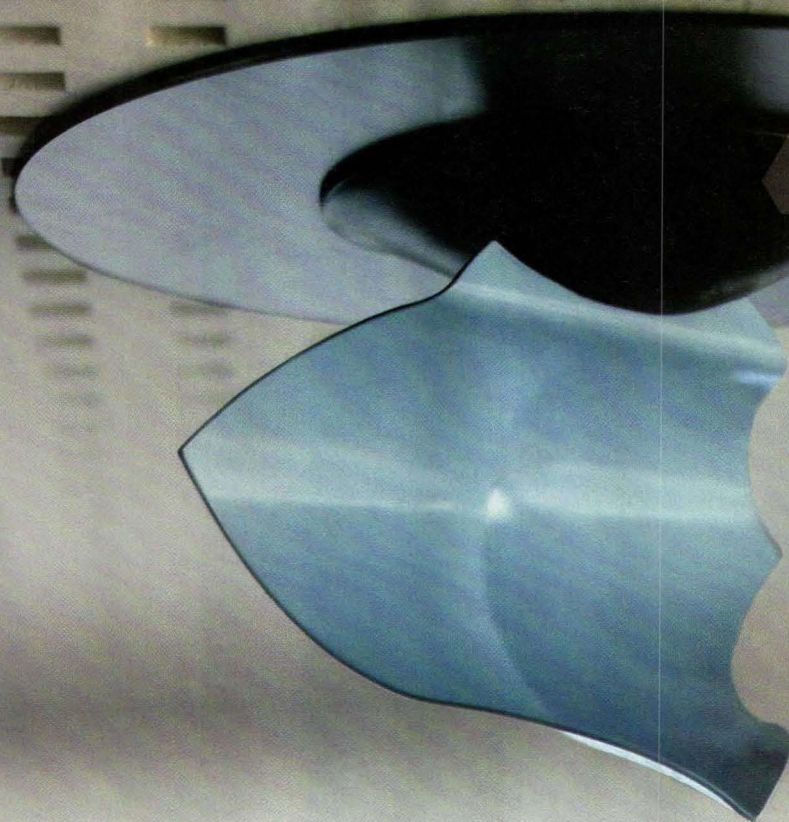


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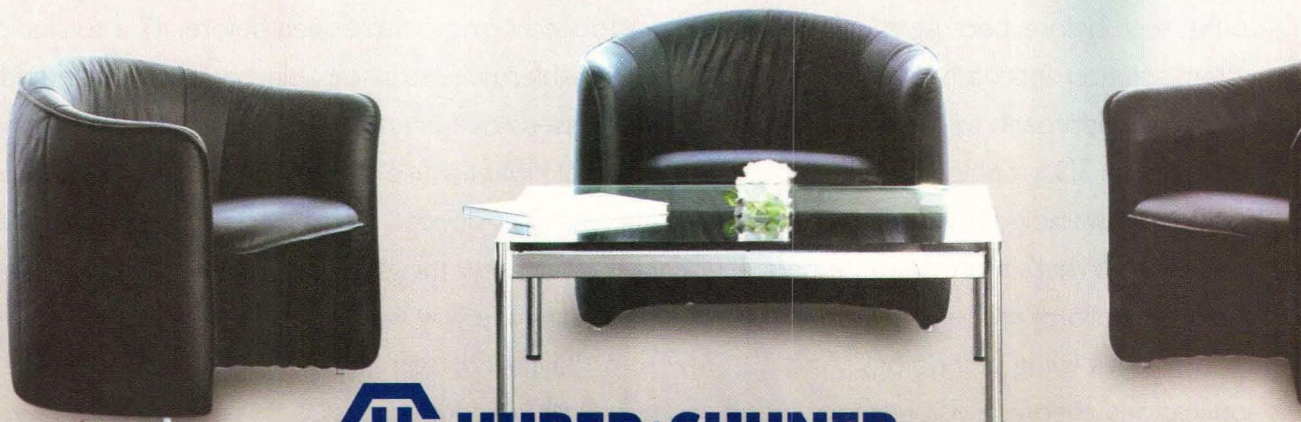
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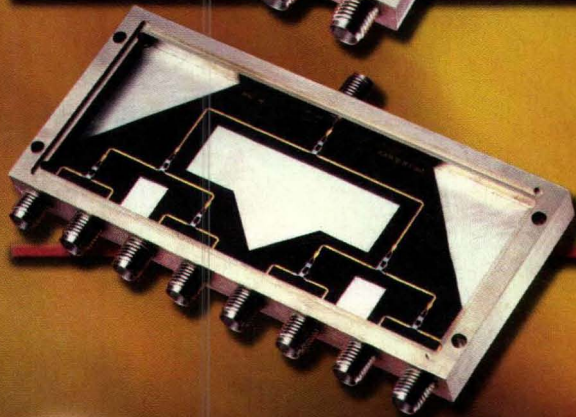
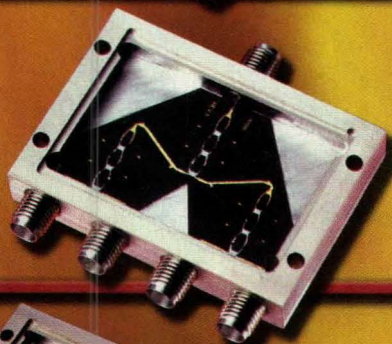
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Phase unbalance	Degrees		± 5.0
Amplitude balance	dB		± 0.5

4 Way Power Divider - Model D0489

RF frequency range	GHz	18	40
Insertion loss	dB		2.5
Isolation	dB	17	
Input VSWR	Ratio		1.8
Output VSWR	Ratio		1.7
Phase unbalance	Degrees		± 5.0
Amplitude balance	dB		± 0.5

8 Way Power Divider - Model D0889

RF frequency range	GHz	18	40
Insertion loss	dB		3.5
Isolation	dB	17	
Input VSWR	Ratio		1.8
Output VSWR	Ratio		1.7
Phase unbalance	Degrees		± 5.0
Amplitude balance	dB		± 0.5

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2003 Wireless Show

►►THE WIRELESS SYSTEMS Design Conference & Expo was a great forum to introduce TRU Draw, our new web-based drawing tool. It was well received by those who stopped by the TRU booth for a demonstration. While the quantity of leads was fewer than hoped for, we did reach a number of high-quality attendees who were mostly design engineers.

Douglas E. Snader

Vice President of Sales and Marketing
TRU Corp.

►►CHIPCON WAS VERY satisfied with the attendance at this year's wireless show, which was held in February in San Jose. We were surprised by the high number of visitors due to the limited number of exhibitors compared to last year's show.

Chipcon also noticed an increased

interest in short-range, low-power, low-data-rate RF ICs from the visitors. Further, we were satisfied with the number of visitors at the half-day technical seminar held by Chipcon during this event. We will hold a similar technical seminar at the MTT show in Philadelphia in June and most likely at next year's wireless show, which we will most definitely attend.

At the wireless show, Chipcon presented its high-performance SmartRF02 product line. The Smart RF02 technology is based on a low-cost, system-on-a-chip (SoC) 0.35- μ m CMOS technology. Each of the four RF ICs that Chipcon has launched from this platform has set new industry standards. The CC1000 RF transceiver offers the short-range industry's best combination of low power consumption, price, and flexibility. The CC1050 is the 300-to-1000-MHz range's lowest-cost and lowest-power-consumption multi-

channel RF transmitter. The CC1010 is the short-range industry's first true RF SoC solution, integrating an RF transceiver and an 8051 microcontroller core on the same die. The latest product release, the CC1020, is the only true narrowband and multichannel RF transceiver below 1 GHz.

Birgit Opland
Chipcon AS



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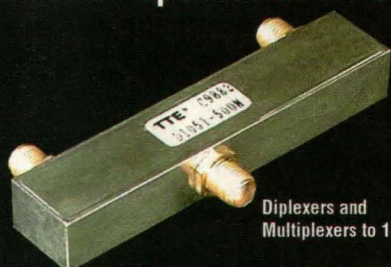
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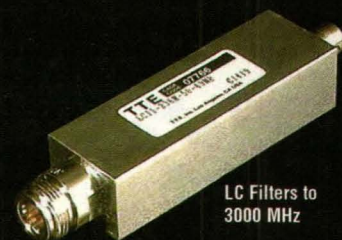
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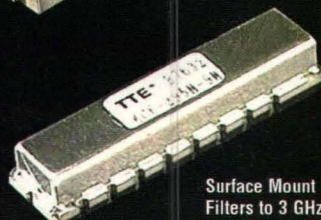
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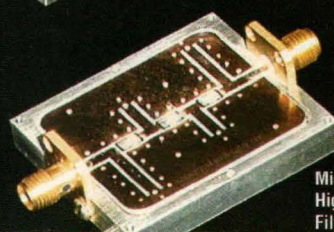
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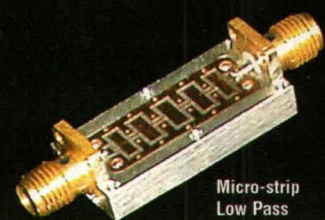
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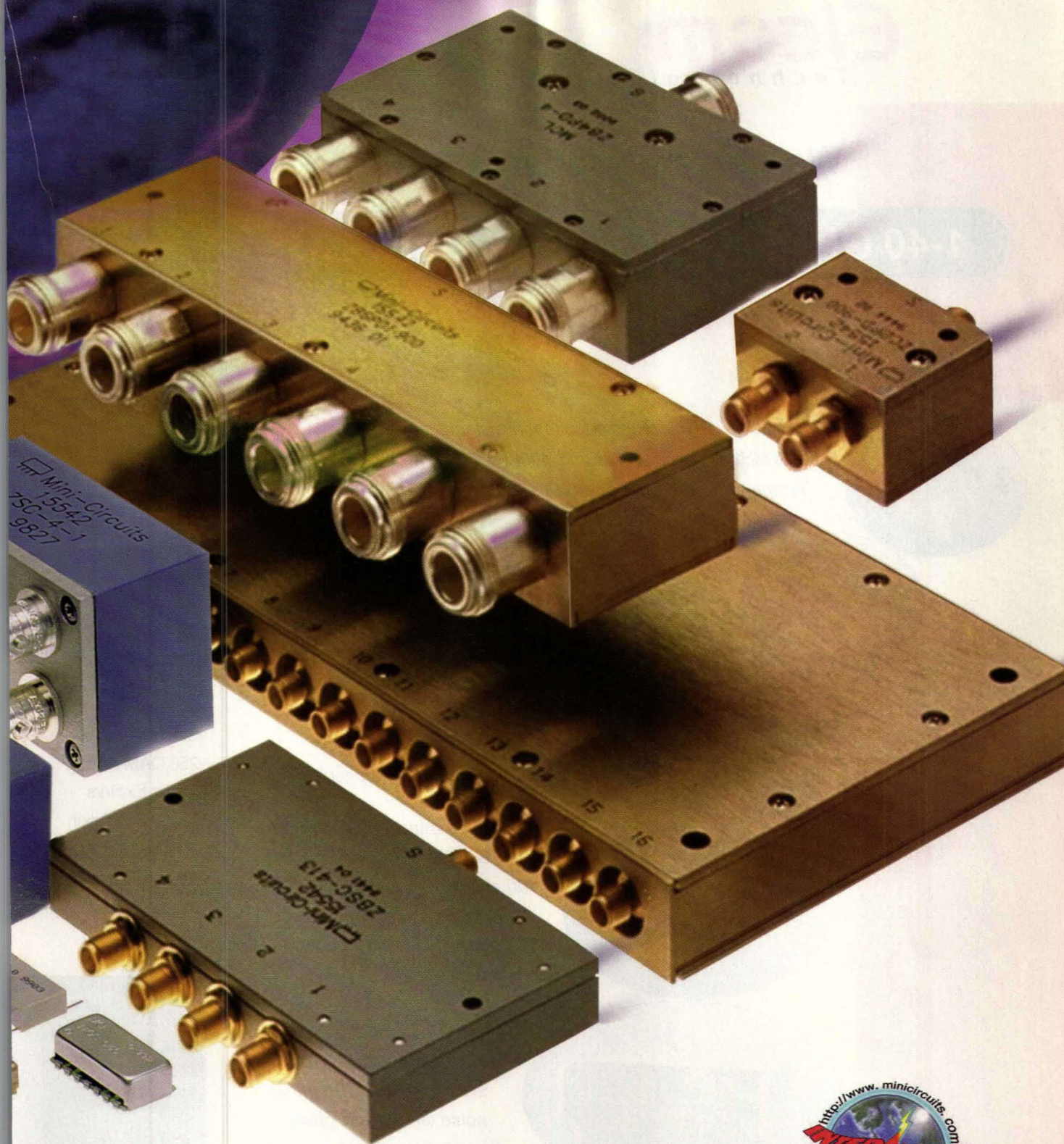
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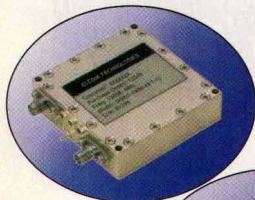
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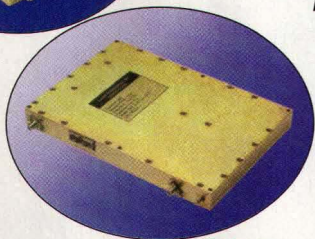
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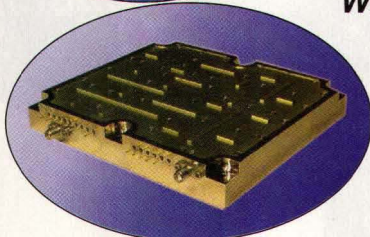
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Looking For A Few Good Presenters

THIS TIME OF YEAR usually evokes images of travel to some distant town to attend the microwave industry's largest technical conference and exhibition, the Microwave Theory & Techniques Symposium (MTT-S). Sponsored by the IEEE, this long-running event, which is also known as the International Microwave Symposium (IMS), is scheduled for June 8-13, 2003 at the Philadelphia Convention Center (Philadelphia, PA). To commemorate the location, this year's meeting has been dubbed "Liberty Through Microwaves," and promises a full slate of technical sessions, tutorial sessions, and workshops. (For a brief preview of the technical sessions and products to be announced at MTT-S, please turn to p. 33).

At one time, the MTT-S was considered the premier event for presenting applications and technology information intended for military use. Although many of the presentations in the 1970s and 1980s tended to cover generic technology issues, the bandwidths of interest (for example, 2 to 18 GHz) generally indicated the type of application that inspired the technology.

With the 1990s, however, came the "age of wireless," and a refocusing of the technology presentations toward more narrowband devices, components, and test equipment. Presentations shifted from discussions of instantaneous operation from 2 to 18 GHz to operation in more specific frequency bands, such as 1.9 and 2.4 GHz. While many of the presentations at this year's MTT-S event are to be lauded for covering important fundamental technologies, such as superconductors and microelectromechanical systems (MEMS), that can be applied to either commercial or military systems, a glance at the program reveals that fewer and fewer presentations are aimed at that important (and growing) engineering community working on the wide-band solutions favored by the military customer.

For that reason, the Military Electronics Show (MES) was started almost three years ago. In part, the event was created out of a need voiced by attendees to the many "commercial-off-the-shelf" (COTS) events who found that they were not learning about the latest significant developments in true high-reliability, ruggedized, military design.

But any event is only as good as the people who support it. Now in its third year, and scheduled for September 16-17, 2003 in the Baltimore Convention Center (Baltimore, MD), the MES is looking for a few good presenters. A full list of proposed topics (such as power supplies, computers and peripherals, receiver design, software, and test) can be found at the website at www.mes2003.com. And if you'd like to be part of a growing event devoted to military design, please drop me a note at jbrowne@penton.com.

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Group Publisher Craig Roth, (201) 845-2448 • croth@penton.com
Publisher/Editor Jack Browne, (201) 845-2405 • jbrowne@penton.com
Technology Editor Nancy Konish, (201) 845-2428 • nkonish@penton.com
Managing Editor John Curley, (201) 845-2415 • jcurley@penton.com
Special Projects Editor Alan ("Pete") Conrad
Editorial Assistant Dawn Prior • dprior@penton.com
Contributing Editors Andrew Laundrie, Allen Podell

MANUFACTURING GROUP

Director Of Manufacturing Ilene Weiner
Group Production Director Mike McCabe
Customer Service Representative
Dorothy Sowa, (201) 845-2453, fax: (201) 845-2494
Production Coordinator Judy Osborn, (201) 845-2445

ART DEPARTMENT

Art Director Armand Veneziano • aveneziano@penton.com
Group Design Manager Anthony Vitolo • tvitolo@penton.com

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EDITORIAL OFFICE

Penton Media, Inc.
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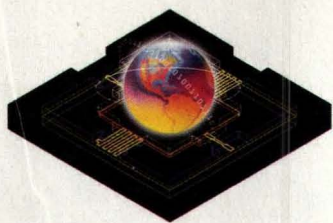
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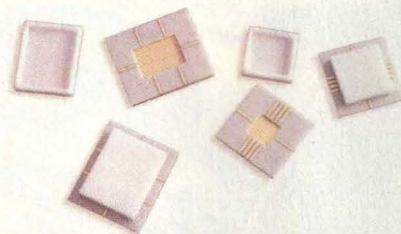
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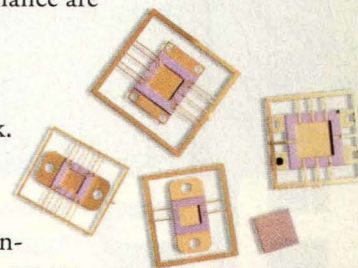
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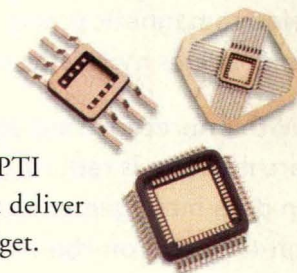
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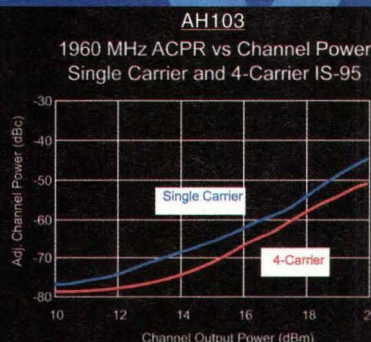
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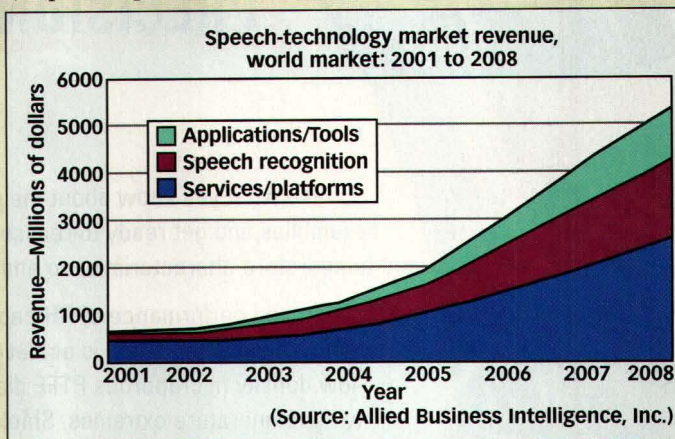
News items from the communications arena.

Speech-Recognition Market To Exceed \$5 Billion By 2008

OYSTER BAY, NY—Speech recognition has slowly been building its reputation, as accuracy rates are much improved and applications to meet users' needs are being developed. An Allied Business Intelligence (ABI) report, "Speech Recognition Systems: Market Opportunities and Major Player Assessment," projects this market to increase to \$897.8 million in 2003, up from \$677 million in 2002. Over the longer term, the speech-recognition market is forecasted to grow to \$5.3 billion by 2008 (see figure).

Interactive Voice Response (IVR) systems based on touchtone input represent a large opportunity for call-center operations. Speech-enabled solutions are making their way as replacement systems for call-center applications. Telecommunications carriers are looking at speech applications as a hook to keep customers from switching services. As the technology for speech becomes available in a smaller footprint, embedded solutions will power handheld computers and wireless phones.

"At the present time, IT budgets are strained and the commitment to develop speech solutions is a large one," explains Edward A. Rerisi, director of research at ABI. "Demonstration of a positive ROI will entice companies to at least get a toe in the water. As users become more aware of the technology and recognize the benefits, demand will help drive applications in carrier and embedded markets."



Electronics Industry Pioneer Harold H. Powell Dies At 84

PHILADELPHIA, PA—Harold H. Powell, founder and chairman of the board of Powell Electronics, Inc., died Friday, March 21, in Miami, FL. Headquartered in Powell's hometown of Philadelphia, Powell Electronics is one of America's largest independent, privately owned electronics distributors.

Powell Electronics' president Ernie Schilling, a 37-year veteran of the company, comments, "Our industry has lost a good and great man. Harold's principles and actions exemplified the way we wish all businesses were run. Those who knew him know just how ethical Harold was and what a true innovator he was. We are committed to maintaining the independence, integrity, and innovation he so cherished."

Powell Electronics, Inc. began operations in 1946 from Powell's apartment on Philadelphia's Spring Garden Street. Powell had purchased 150 surplus microamp meters for \$100 and their rapid resale at \$2.45 each was the foundation on which the company was built. Shortly after the company's founding, it moved to a storefront on Arch Street in Philadelphia.

In 1954, Bendix appointed Powell Electronics as one of its first authorized distributors. By 1964, the company opened its first branches in Beltsville, MD and Fayetteville, TN. The following year, the company expanded to Florida and the West Coast. Currently, there are 12 branches throughout the country and 200 employees.

Powell retired from the presidency of the company in 1978 and moved up to chairman of the board, a position he held until his death.

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Broadband Needs Clear Regulatory Vision

CAMBRIDGE, ENGLAND—The economies of Europe and the US might miss out on the full benefits of broadband unless an appropriate regulatory vision is created for the future of the telecommunications industry, according to a report entitled, *Regulating the Telecoms Market: competition and innovation in the broadband economy*. The report was released by Analysys, a global adviser on telecommunications, IT, and media.

"The advent of broadband technologies has enabled the beginnings of a fully 'networked economy,' where people and IT systems can communicate ubiquitously, rapidly, and cheaply through always-on fast communication networks," says Ross Pow, one of the authors of the report.

"We have already seen some big changes to the way that companies do business, to where people work, and to how governments interact with their citizens, and broadband offers the basis for a huge amount of innovation across the whole economy, much of which cannot be predicted at the moment," continues Pow.

However, the report highlights that the structure of the broadband market means that realizing that this innovation and its associated benefits will not be straightforward. The deployment and delivery of broadband involves a number of different layers (ducts, cables, network transmission, services, and content and applications), each of which has quite different economies.

"While competition will be the primary means by which economic benefits will be secured, with lots of players offering innovative services and applications," adds Pow, "it may be necessary to consider a market structure that has fewer players at the duct and cable layers in order to realize the required level of investment in rapid, widespread, and commercially viable broadband rollout."

This, according to Analysys, requires regulators and policymakers to consider continuing with an industry-specific regulatory model, rather than moving to a purely competition-based approach as it is currently being pursued. Acceptance that is appropriate for the regulation of broadband telecoms to take into account the potential scale of benefits that it can bring to the wider economy will also be required.

The report recommends that a primary goal for regulators and policymakers, in developing a vision of the telecoms market, should be to maximize the level of innovation in networks, services, and applications and, while competition at the higher layers is essential in achieving this, the heavy investment in the underlying physical infrastructure may require alternative approaches to ensuring the ubiquitous availability of broadband access.

Agere Enters Wireless Power-Amplifier Market

ALLENTOWN, PA—Agere Systems has unveiled 21 transistors targeting the wireless base-station power-amplifier (PA) market. Agere's products are targeted for third-generation (3G), 2.5 generation (2.5G), and second-generation (2G) base-station equipment.

Agere's PA transistors can enable much cooler, smaller, and less-expensive wireless base stations than are possible using any other RF power-transistor technology.

In addition, Agere's products help accelerate the industry trend to shrink the size and shift the location of today's typical base stations, about the size of a backyard toolshed and installed on the ground, to the size of a suitcase and installed above the ground on wireless antenna towers.

With these products, Agere is the first to achieve the transistor temperature (thermal) performance level that the industry has been striving to attain for the past 10 years. The transistors achieve 10-to-15-percent lower operating temperatures than all of the other competing transistors available today.

Agere's lower temperature transistors can cut in half the number of cooling fans in base stations compared with hotter transistor products. Reducing the number of fans also reduces noise pollution, a major issue in the base-station market.

"The wireless transistor market represents an important new growth opportunity for Agere, and with our technological breakthroughs, we believe we are poised for success in this space," comments Sohail Khan, executive vice president of Agere's Infrastructure Systems Group. "By delivering significant cost reductions, our new products will enable wireless service providers to accelerate delivery of lower-cost, feature-rich, high-data-rate services to cell-phone users, such as video streaming, instant messaging, and gaming."

“The deployment and delivery of broadband involves a number of different layers, each of which has quite different economics.”



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Mini-Circuits VAT and HAT fixed attenuators rank at the top of their class for high performance, big selection, and low cost! Choose from 14 different attenuation values; from 1 to 10dB in 1dB steps plus 12, 15, 20, and 30dB. All in stock, ready for immediate shipment, and *value priced* from only \$9.95 for BNC (HAT) and \$11.95 for SMA (VAT). Performance wise, these devices offer excellent attenuation flatness, low VSWR, and handle up to 500mW input power. Plus, rugged unibody construction makes them very easy to use in systems, testing, and product development applications. So get the best economy from your design with Mini-Circuits fixed attenuators.

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Models		Attenuation* (dB)			VSWR (:1)	
SMA-M/F	BNC-M/F	Nominal	Flatness		Midband Typ.	Midband Typ.
DC-6GHz	DC-2GHz		Midband	Typ.		
VAT-1	HAT-1	1	1	0.20 0.11	1.10	1.2
VAT-2	HAT-2	2	2	0.20 0.10	1.20	1.2
VAT-3	HAT-3	3	3	0.15 0.12	1.15	1.1
VAT-4	HAT-4	4	4	0.15 0.08	1.15	1.1
VAT-5	HAT-5	5	5	0.10 0.06	1.15	1.1
VAT-6	HAT-6	6	6	0.10 0.02	1.15	1.1
VAT-7	HAT-7	7	7	0.10 0.05	1.15	1.1
VAT-8	HAT-8	8	8	0.10 0.04	1.20	1.1
VAT-9	HAT-9	9	9	0.10 0.02	1.15	1.1
VAT-10	HAT-10	10	10	0.20 0.03	1.20	1.1
VAT-12	HAT-12	12	12	0.10 0.05	1.20	1.1
VAT-15	HAT-15	15	15	0.30 0.05	1.40	1.1
VAT-20	HAT-20	20	20	0.75 0.18	1.20	1.1
VAT-30	HAT-30	30	30	0.30 0.38	1.15	1.1

Power: 0.5W at 70°C ambient.

* Attenuation varies by ± 0.3 dB max. (VAT), ± 0.2 dB max. (HAT) over temperature.

• VAT MODELS \$11.95 ea. (qty. 1-9) • HAT MODELS \$9.95 ea. (qty. 1-9)

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K3-VAT: 2 of Ea. VAT-3, -6, -10 (6 total) \$59.95

K1-HAT: 1 of Ea. HAT-3, -6, -10, -20, -30 (5 total) \$48.95

K2-HAT: 1 of Ea. HAT-1, -2, -3, -4, -5, -6, -7, -8, -9, -10 (10 total) \$97.95

K3-HAT: 2 of Ea. HAT-3, -6, -10 (6 total) \$58.95

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Agilent's EDA Division Joins Georgia's Broadband Initiative

ATLANTA, GA—Yamacraw, the state's economic-development initiative that is helping Georgia become a leader in broadband-communications-technology design, has announced that a division of Agilent Technologies, Agilent EESof EDA, a supplier of electronic-design-automation (EDA) software for broadband communications, has joined as a full member. Through its membership, Agilent will provide software and hardware to help create state-of-the-art electronic capabilities in Georgia's research universities. "We're pleased to join the growing list of Yamacraw members," says Charles Plott, product marketing manager at Agilent EESof EDA. "Our membership connects us with Georgia's vast pool of technical talent, and also with our customers in the communications industry throughout the southeast."

"Agilent's EESof EDA is helping build tomorrow's high-tech communications products today," states Herb Lehman, Yamacraw director. "Through cutting-edge research and product development, they are changing the face of communications. Their research and capabilities will contribute greatly to Georgia's efforts and is an ideal fit for Yamacraw."

Agilent will benefit from its investment with Yamacraw by ensuring that Georgia's university students are trained and experienced on their tools and technology, according to Plott.

Kudos

CLEVELAND, OH—The American Physical Society (APS) named Arthur Ashkin of Bell Laboratories as the 2003 winner of the Joseph F. Keithley Award. The award was given for Ashkin's theoretical and experimental contributions to the understanding of laser cooling and trapping of atoms and particles, for demonstrating the optical gradient forces on atoms and the trapping of atoms with light, and for inventing optical tweezers and showing how they can be used to measure the physical forces generated by biological molecular motors.

The Keithley Award, established in 1997, presents \$5000 annually to a physicist who has been instrumental in the development of measurement techniques or equipment that have an impact on the physics community by providing better measurements. The award

honors Joseph F. Keithley, founder of Keithley Instruments, Inc., for his contributions in the area of sensitive and precision instrument development and measurement techniques.

GREENSBORO, NC—RF Micro Devices, Inc., a provider of proprietary RF integrated circuits (RF ICs) for wireless-communications applications, announced that it has received ISO 14001:1996 certification for environmental management systems by the TUV Management Service certification body.

Certification was granted upon independent verification of RFMD's compliance with the standard.

SAN JOSE, CA—Parthus Ceva, Inc., a provider of licensable digital-signal-processor (DSP) cores and platform-level IP solutions to the semiconductor industry, announced an extensive upgrade to PLLXpert Online with the addition of the TSMC 0.13G and TSMC 0.13LV processes, 'ready to go' PLLs, and strong industry adoption of PLLXpert Online.

Launched in 2002, PLLXpert Online (www.pllxpert.com) has captured over 25 new licensing customers in the past year since launch. The latest upgrade to PLLXpert Online adds multiple new process options on TSMC, UMC, Silterra, and First Silicon (spanning 0.25-, 0.18-, and 0.13- μ m geometries), and a portfolio of 'ready-to-go' application-specific PLLs available for download.

VISTA, CA—Palomar Technologies, Inc., a manufacturer of assembly systems, announced that its Gold Connection™ Flip Chip Bonding System has been selected as a *Fiberoptic Product News* 2002 Silver Technology Award Winner in the category of Automation/Packaging.

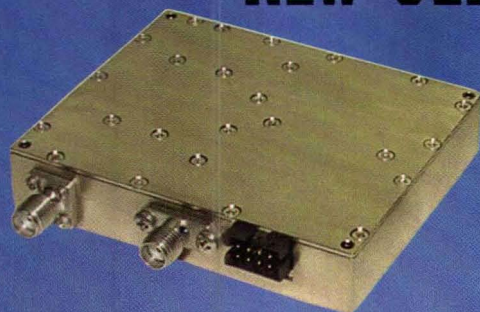
Winning products are selected based on technological innovation, design, usefulness, and response from *Fiberoptic Product News* (FPN) readers and editors. FPN, a Reed Business Publication, covers fiber-optic technology applications. The publication has 37,000 readers.

Winners were selected from the more than 2000 new products that appeared in FPN in 2002. After a vote by the 37,000 FPN readers, the top 50 were reviewed by FPN editors, who chose 24 nominees in six product categories. Final voting took place via the Internet on the FPN website. NEENAH, WI—Plexus Corp. announced that it has received the 2003 Overall Service Excellence Award in the large-company category (annual revenues over \$500 million) for Electronics Manufacturing Services (EMS) providers, sponsored by *Circuits Assembly* magazine. **MRF**

“Agilent EESof EDA is helping build tomorrow's high-tech communications products today.”

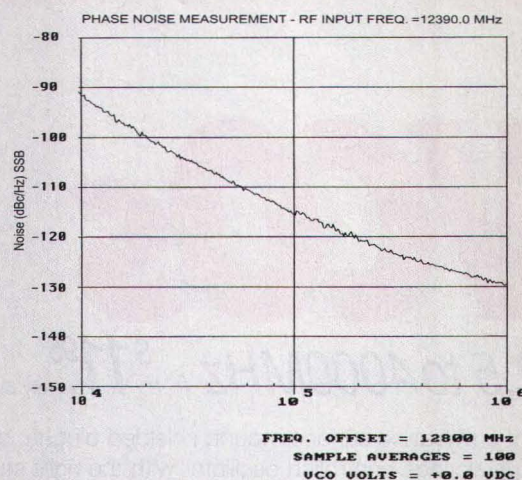
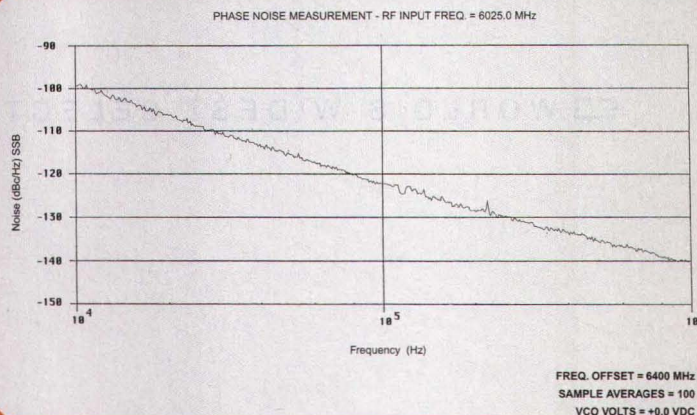
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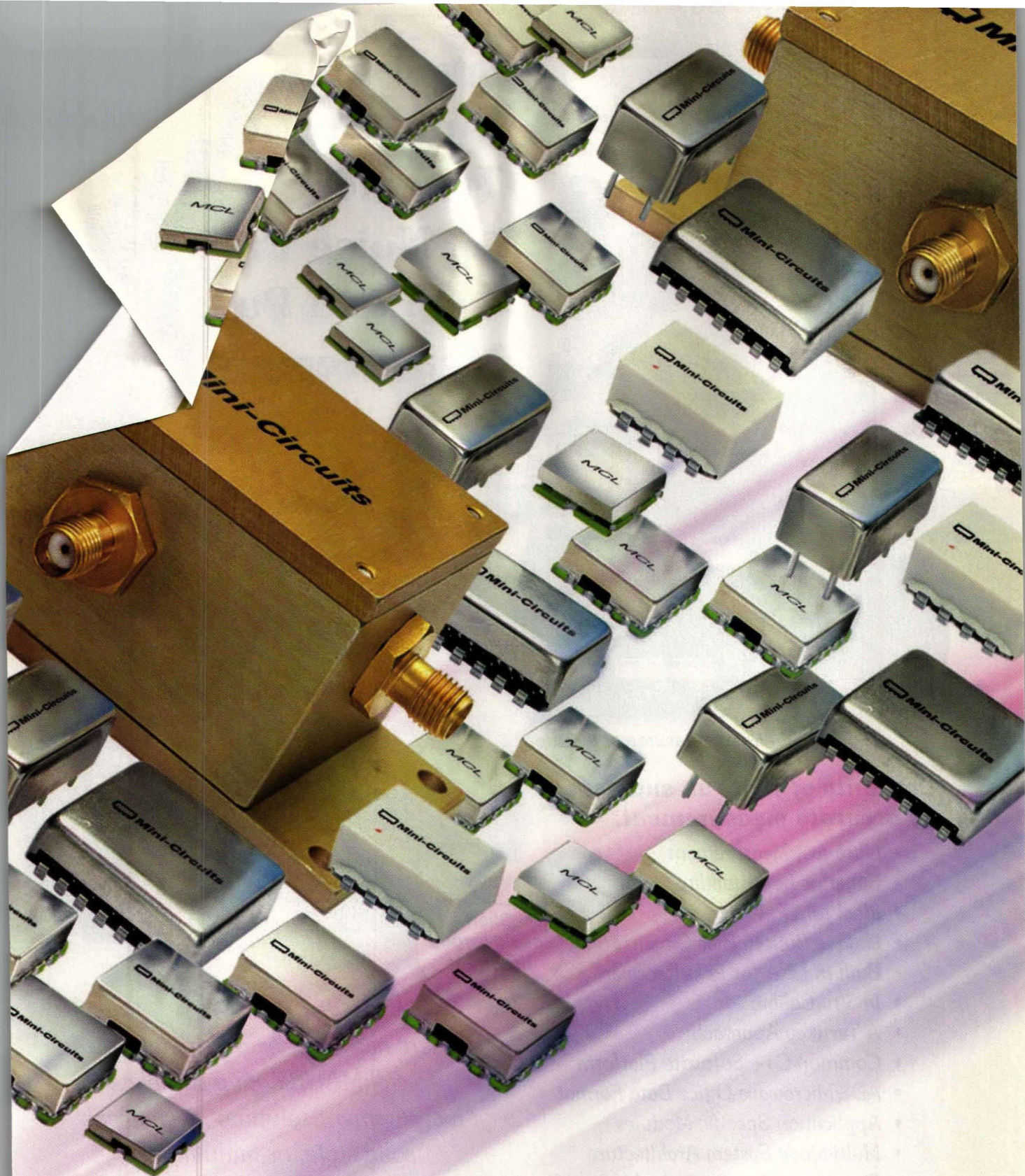
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Show Summarizes Microwave Technology

This year's edition of the International Microwave Symposium, which will take place in Philadelphia, promises a full schedule of workshops, technical sessions, and product exhibitions.

microwave engineers usually mark their calendars for at least one key event each year: the IEEE's Microwave Theory & Techniques Symposium (MTT-S). Also known as the International Microwave Symposium (IMS), the high-frequency meeting is scheduled for June 8-13, 2003 in the Philadelphia Convention Center (Philadelphia, PA). As part of a full week of technical sessions, the event also includes a three-day

Radio-Frequency Integrated Circuit (RF IC) Symposium (June 8-10) and a two-day Automatic RF Techniques Group (ARFTG) meeting (June 12-13) devoted to improved microwave measurement and calibration methods.

The theme of this year's event is "Liberty Through Microwaves." The IMS technical sessions run from June 10th through June 12th, with workshops and tutorial sessions scheduled for June 8th, 9th, and 13th. Tuesday's (June 10th) technical sessions include discussions on linearization techniques for high-power amplifiers, advances in high-power transistor technologies, novel modeling and computer-aided-design (CAD) techniques (including the use of neural networks and fuzzy logic), millimeter-wave monolithic IC technologies, microwave photonic devices, and the biological effects of microwaves and medical applications for RF technology.

Wednesday's (June 11th) technical sessions is strong on presentations for passive-component researchers, includ-

ing discussions on planar technologies for filters and multiplexers, novel waveguide structures, leakage

effects in planar structures, and techniques for designing cavity filters and multiplexers. Additional sessions cover nonlinear device modeling techniques, power amplifiers (PAs) and devices for wireless applications, behavioral modeling for nonlinear devices and components, millimeter-wave signal generation and amplification software radios, Terahertz technologies, ICs for optical communications, and time-domain techniques.

Finally, Thursday wraps up the main technical sessions with several presentations of interest for those studying microelectromechanical systems (MEMS). For example, separate sessions explore MEMS techniques for tunable filters and resonators, future material technologies for RF MEMS, RF MEMS phase shifters and micromachined inductors, and modeling and packaging of RF MEMS components. Additional technical sessions detail microwave filter-synthesis techniques, superconducting filters, microwave and millimeter-wave sensor applications, low-noise compo-

JACK BROWNE
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Managing Editor

nents based on silicon (Si), gallium-arsenide (GaAs), and indium-phosphide (InP) technologies, advances in beam-steering and beam-forming arrays, spatial power-combining techniques, advanced millimeter-wave transceiver and source technology, ferroelectric

and acoustic devices, and high-speed sampling circuits and techniques.

Philadelphia's MTT-S event features several workshops (on microwave filter design and microwave oscillator design) and a wide array of workshops. For example, Ferdo Ivanek has organized

a Monday workshop on fixed broadband wireless applications entitled "Bridging the Last Mile: Technology Push Versus Market Pull in Fixed Broadband Wireless Access (BWA)," while that same day Sam Horowitz of Dupont (Wilmington, DE) offers views on ceramic packaging technologies with "Latest Advances in Ceramic Interconnect Technologies." Sunday workshops include an update on ultrawideband technology, "Ultrawideband: Theory and Implementation" by David Lovelace, "Recent Developments in Oscillator Design," organized by Steve Maas of Applied Wave Research (El Segundo, CA), and "Next Generation Transmitter Architecture and Design," organized by Ed Niehenke, formerly of Westinghouse (Baltimore, MD). More information on the MTT-S meeting in Philadelphia is available by visiting the IEEE's website at www.IEEE.org. In addition, what follows is a brief look at some of the new products expected to be on display from key MTT-S exhibitors.



Winchester Electronics

BMA connectors are designed for use in blindmate applications requiring multiple connectors to be mated and unmated simultaneously. This blindmate capability is accomplished by the slide-on, non-locking BMA interface - designed to allow up to .020" radial and .060" axial misalignment. Additional features include:

- 50 Ohm
- Frequency range: 0 - 18 GHz
- Low mating force (3.0 lbs.) provides easy mating and unmating
- Hot Swappable - ground contact engages before signal
- Vertical and right-angle mount PCB plugs and jacks
- Bulkhead and snap-in mount plugs and jacks
- 2-hole and 4-hole panel mount plugs and jacks
- Flexible and semi-rigid cables
- Custom cable assemblies designed to your specifications

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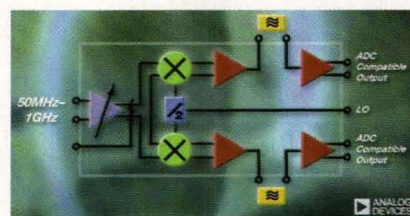
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Wallingford CT 06492
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Demodulator Enables IF-To-Baseband Conversion From 50 MHz To 1 GHz



THE AD8348 HIGH-PERFORMANCE broadband quadrature demodulator is designed to enable IF-to-baseband conversion from 50 MHz to 1 GHz. In addition to the demodulator, which includes a dual mixer core conversion and local-oscillator (LO) phase splitter, the AD8348 integrates a 45-dB linear-in-dB variable-gain amplifier (VGA) and single-ended to differential amplifiers for driving baseband ADCs. The AD8348 is optimized and suited for driving dual-channel, low-cost CMOS ADCs, such as ADI's AD9218.

With high linearity, good amplitude/phase balance, and 60-MHz demodulation bandwidth, the AD8348 enables most high-order modulation

SUPER FAST VERY HIGH ISOLATION SWITCHES



\$195*
SPDT, DC-5GHz From ea. (10,000) **IN STOCK**

Very high isolation up to 90dB at 1GHz typical. Built-in TTL driver with blazing fast 10nsec switching speed. The ability to withstand severe operating temperatures down to -40°C to +85°C. That's what's great about Mini-Circuits wideband surface mount and coaxial SPDT switches. But that's not all! Reflective and absorptive models are available in three different package styles to suite your design requirements; M3SW's 3x3mm MCLP™ surface mount package with exposed metal bottom for excellent grounding and heat dissipation, SWM's SOIC-8 for easier assembly, and ZASW's tough built coaxial design with SMA-F connectors. No matter which model you choose, you'll get strong performance and rugged reliability at a price that crushes the competition. Check all the specs on our web site, then contact Mini-Circuits for our fast response!

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SPECIFICATIONS (@ 1GHz)

Model	Freq. (GHz)	In-Out Isol. dB(typ)	Ins. Loss dB(typ)	1dB Comp. dBm(typ)	Price \$ea. (Qty. 10)
• M3SW-2-50DR	DC-4.5	60	0.7	25	4.95 *
■ M3SWA-2-50DR	DC-4.5	65	0.7	25	4.95 *
• SWM-2-50DR	DC-4.5	55	0.7	25	5.30
■ SWMA-2-50DR	DC-4.5	65	0.7	25	5.30
• ZASW-2-50DR	DC-5	90	1.7	20	79.95
■ ZASWA-2-50DR	DC-5	90	1.7	20	79.95

Supply voltage +5V, -5V, TTL control.
Switching time 10nsec (typ).
• Reflective ■ Absorptive



ZASW-2-50DR
ZASWA-2-50DR

ACTUAL SIZE



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formats, including QAM, QPSK, and 8-PSK. This combination of performance, broadband operating frequency, and flexible device architecture makes the AD8348 suitable for a variety of wireless networking applications, including cellular infrastructure CDMA/WCDMA/GSM EDGE and high-capacity, point-to-point and point-to-multipoint radio links, WLAN, and wireless-local-loop equipment.

P&A: The AD8348 is available in a 28-lead TSSOP and is fully specified for use over the -40°C to $+85^{\circ}\text{C}$ temperature range. Samples are pre-production quantities are available for immediate shipment. The AD8348 is priced at \$4.95 per unit in 10,000-piece quantities.

Analog Devices, Inc., 804 Woburn St., Wilmington, MA 01887; (800) 262-5643, FAX: (781) 937-1021, Internet: www.analog.com.

MMIC Amplifier Die Cover DC To 10.0 GHz

A FAMILY OF FOUR InGaP HBT Gain Block MMIC amplifiers covers DC to 10.0 GHz. These amplifier die can be used as either cascadable 50- Ω gain stages or to drive the LO of HMC mixers with up to 17-dBm output power, making them a suitable choice for Microwave P2P/VSAT, test-equipment, Military EW/ECM/C31, and space-telecommunications applications.

Both the HMC395 and HMC405 offer 16 dB of gain with output IP3s of +31 dBm and +32 dBm, respectively. The HMC396 provides 12 dBm of gain, has output IP3 of +30 dBm, and covers applications in the DC-to-8.0-GHz band. The HMC397, with 15 dB of gain and output IP3 of +32 dBm, covers the DC to 10.0-GHz market. All products require only 50 to 56 mA from a +5-V supply.

The family of Gain Blocks utilizes a Darlington feedback pair which results in reduced sensitivity to normal process variations and yields excellent gain stability over temperature while requiring a minimal number of external components. All of these MMICs can be easily integrated into Multi-Chip-Modules

(MCMs) due to their small size.

P&A: Sample and production quantities are available.

Hittite Microwave Corp., 12 Elizabeth Dr., Chelmsford, MA 01824; (978) 250-3343, FAX: (978) 250-3373, Internet: www.hittite.com.

Snap-On Connector Is Easier And Faster To Mate

THE QN QUICK-LOCK N SIZE connector requires only a very low mating force of about 30 N and is up to 10 times faster to mate than N type connectors, as the locking mechanism snaps closed in a



single step. Polling back the de-coupling sleeve opens the snap ring of the mating mechanism, allowing for quick de-mating with a force of merely 30 N.

The quick mating cycle of the QN connector leads to a much reduced cost of ownership compared with series N connectors, yet it delivers the same performance. The device can be used in a wide variety of applications such as radio base stations, antenna systems, and test and measurement, primarily all applications where medium or high power has to be transmitted with low loss.

The QN connector is rugged and water tight and features good inter-modulation characteristics, low return loss, and suitable RF leakage. Due to the reduced flange size (like TNC) of the panel connectors, a higher packing density is achieved. QN angle connectors may also be aligned to the desired position after having coupled them with the counterpart. Customized solutions for the QN Quick-Lock connector are also available.

HUBER+SUHNER AG, Mobile Communica-

tions + Electronics, 9100 Herisau, Switzerland; +41 (0)71 353 41 11, FAX: +41 (0)71 353 44 44, Internet: www.hubersuhner.com

Radio Transceiver Integrates Rx And Tx

A SINGLE-CHIP RADIO TRANSCEIVER in a compact package provides the radio receive and transmit functions for radio-communications systems in the unlicensed 2.4-GHz ISM frequency band. Designated TB32301AFL, the low-cost monolithic device also incorporates a low-noise amplifier (LNA), frequency-modulation (FM) detector, voltage-controlled oscillator (VCO), power amplifier (PA), and received strength signal indicator (RSSI). It is targeted at a wide range of RF communication applications, including remote control, office, and building security and wireless home networks.

Specifications include power supply voltage of 2.7 to 3.3 V (with an operating temperature range of -20°C to 70°C), clock frequency of 4 to 20 MHz, and operating frequency of 2400 to 2500 MHz.

P&A: Engineering samples of TB32301AFL are available now and are priced at \$1.99 per piece in 100,000-piece quantities. Mass production is slated to begin in June 2003.

Toshiba America Electronics Components, Inc.; Internet: www.chips.toshiba.com.

Mini-SSPAs Offer Small Size Combined With Good Performance

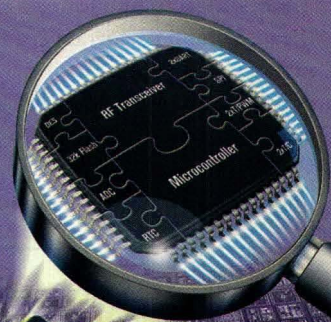
THE SOPHIA WIRELESS MINI-SSPA configurations provide a combination of compact size and linear power performance. The inline waveguide combining and modular construction provides a variety of choices to meet custom RF power requirements and mechanical configurations. Due to the high linearity of solid-state devices, minimal "back-off" from max. power is required for multi-carrier or other linear sensitive applications.

The DC power-management module can be configured to provide voltage regulation, bias sequencing, current monitoring, and thermal monitoring.



- Connecting Smarter

RF Transceivers 300-1000 MHz single-chip, low-power, low-cost CMOS RF-ICs based on the SmartRF™02 technology



CC1000/CC1050

- Very low power RF Transceiver and RF Transmitter
- 315/433/868/915 MHz

CC1010

- Industry's first integrated RF Transceiver and 8051 microcontroller
- 315/433/868/915 MHz

CC1020

- Industry's first true narrow band RF Transceiver
- 402-470/804-940 MHz

Applications

- Home automation
- Automatic meter reading systems
- Remote keyless entry
- Wireless alarm and security systems
- Narrow band applications
- Game controllers and electronic toys
- Wireless headset
- Telemetry systems

Features	CC1000	CC1050	CC1010	CC1020
Product	Transceiver	Transmitter	Transceiver	Transceiver
Programmable frequency	300 - 1000 MHz	300 - 1000 MHz	300 - 1000 MHz	402-470/804-940 MHz
Supply voltage	2.1 - 3.6 V	2.1 - 3.6 V	2.7 - 3.6 V	2.3 - 3.6 V
Current consumption (RX)	7.4 mA	NA	13.1 mA*	16.9 mA
@ 0 dBm (TX)	10.4 mA	9.1 mA	14.4 mA*	13.7 mA
FSK data rate	76.8 kbit/s	76.8 kbit/s	76.8 kbit/s	153.6 kbit/s
Multi channel systems / frequency hopping protocols	✓	✓	✓	✓
RSSI output	✓	NA	✓	✓
Integrated bit synchronizer	✓	NA	✓	✓
Modulation format	FSK/ASK	FSK/ASK	FSK/ASK	FSK/ASK/GFSK
Receiver sensitivity	-110 dBm	NA	-107 dBm	-121 dBm
Programmable output power ranging from	-20 to 10 dBm	-20 to 12 dBm	-20 to 10 dBm	-20 to 10 dBm
Internal RF switch / IF filter	✓	-	✓	✓
Antenna connection	Single ended	Single ended	Single ended	Single ended
Package type	TSSOP-28	TSSOP-24	TQFP-64	QFN-32
Complies with EN 300 220 and FCC CFR 47, part 15	✓	✓	✓	✓
Narrow band (12.5/25 kHz channels)	-	-	-	✓
Integrated microcontroller	-	-	✓ 8051-core	-

* 3.6864 MHz crystal oscillator

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Meet in conference room 303A at 1:00 p.m. on Wednesday 11 June 2003, in the Pennsylvania Convention Center.

For information on Chipcon's complete product range, you are also welcome to visit Chipcon at stand 717.

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Range Of WLAN Antenna Feeder Cable Assemblies Is Available

A WIDE RANGE OF antenna cable assemblies for Cisco/Aironet WLAN bridges and access points is available. The reverse polarity TNC connectors used on these assemblies have been optimized to provide very low VSWR in both the 2.4- (802.11b) and 5.8-GHz (802.11a) band for the highest data rates. These assemblies are available in the standard LMR®



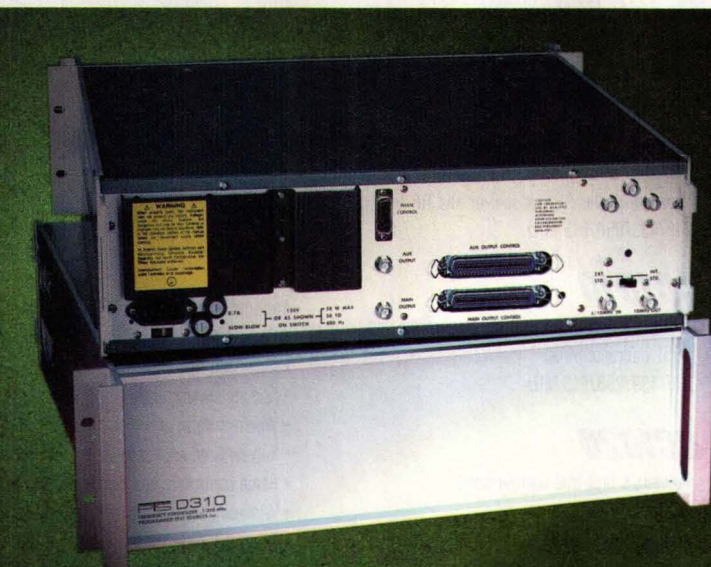
construction for outdoor application as well as other cable constructions for other installation environments.

The Times Microwave cable assemblies are 100-percent tested for VSWR and insertion loss. There is Max. VSWR of 1.25:1 in the 2.4-GHz band and 1.35:1 in the 5.8-GHz band. The assemblies are available in LMR-200, 400, and 600. For longer runs and lower loss, LMR-900-DB assemblies can be provided. **Times Microwave Systems, 358 Hall Ave., Wallingford, CT 06492; (800) 867-2629, (203) 949-8400, FAX: (203) 949-8423, Internet: www.timesmicrowave.com.**

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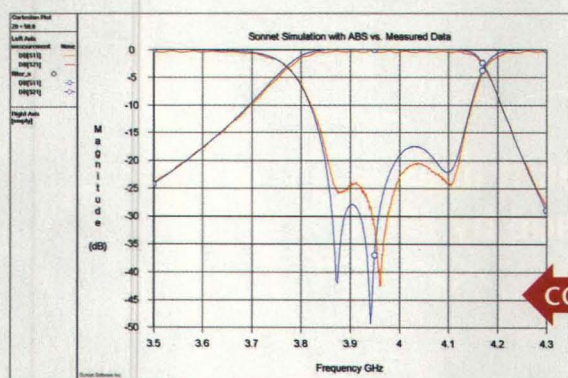
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Adaptive Band Synthesis for 3D Planar EM Simulation

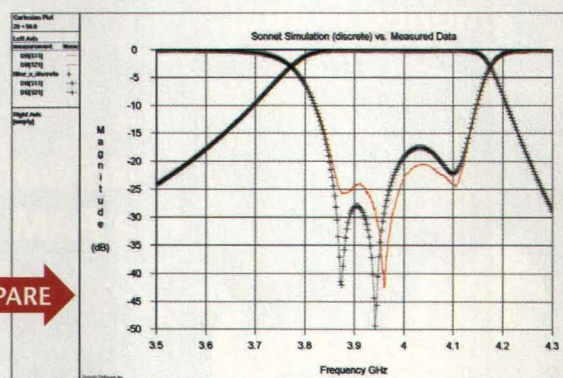
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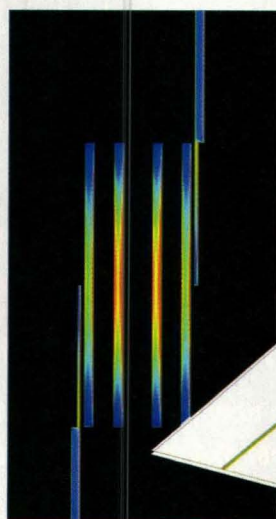
ABS uses the smallest number of discrete EM simulation samples possible, and provides a broadband S-, Y- or Z-parameter data sweep, cutting overall simulation time dramatically and filling in the fine spectral behavior with no reduction in accuracy! And it's reliable and stable for bandwidths exceeding 100x. Compare the results below between measured and calculated using an ABS sweep based on 4 discrete EM analysis frequencies on the left, and a full discrete frequency by frequency simulation on the right.



ABS simulation data based on 4 discrete EM analysis frequencies and measured data



300-point Discrete EM analysis and measured data



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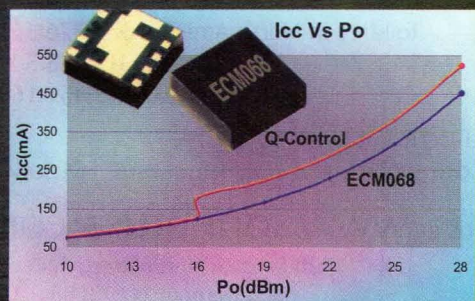
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operating frequency range of 390 to 400 MHz, while providing return loss of 25 dB typical, insertion loss at 0.5 dB maximum, and isolation at 55 dB minimum. This dual-junction unit is available in all TETRA frequencies with integrated reverse power detector, optional high reverse power detector, and optional connector types.

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A HIGH-LINEARITY enhancement-mode pseudomorphic high electron mobility transistor (E-pHEMT) field-effect transistor (FET) is a new release that is designed for low-noise, high-dynamic-range operation in cost-sensitive wireless-infrastructure applications that operate between 450 MHz and 6 GHz.

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The ATF-58143 E-pHEMT FET is housed in the miniature 2.0 × 2.1-mm SOT-343 package.

P&A: The Agilent ATF-58143 is priced at \$1.08 at 10,000-to-24,999-piece quantities.

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MODEL AML0126L2301 IS A low-noise, ultra-broadband amplifier. Operating in the frequency range of 100 MHz to 26.5 GHz, this product provides 23 dB gain and +8 dBm output power at 1 dB gain compression. Input and output VSWR is 2.5:1 nominal. Operating with a voltage of +15 VDC, this amplifier draws a nominal 180 mA.

Internal DC regulator, reverse voltage protection and field-removable SMA (insert m/f) connector shells are standard. This product is packaged in an AML housing that measures 0.99 × 0.75 in. (2.51 × 1.91 cm). Modules are also available as carrier-mount substrates.

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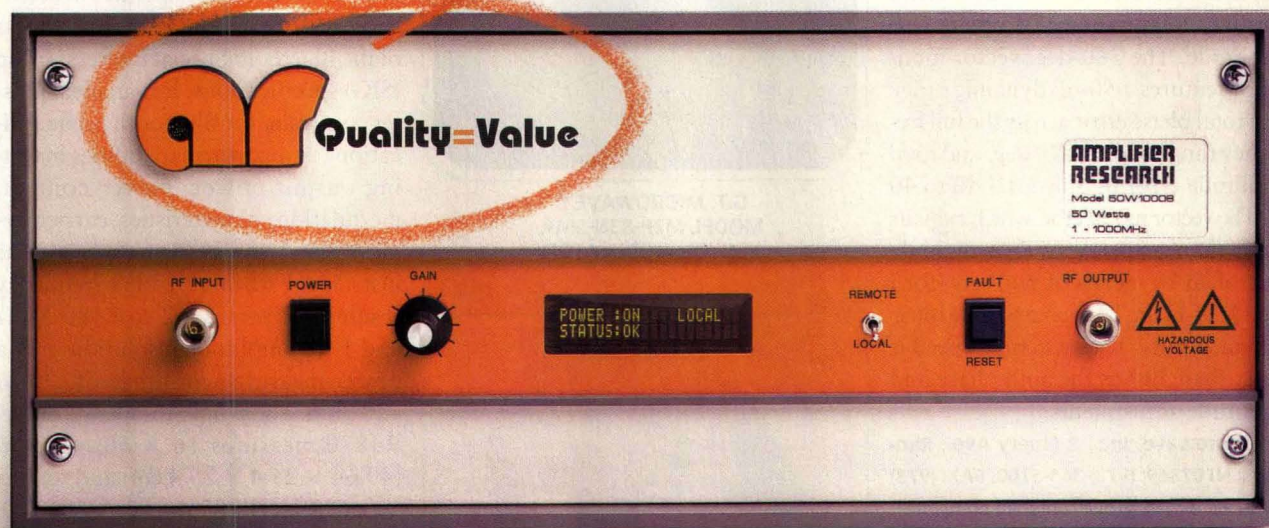
Ansoft Corp., Four Station Square, Suite 200, Pittsburgh, PA 15219-1119; (412) 261-3200, FAX: (412) 471-9427, e-mail: info@ansoft.com, Internet: www.ansoft.com.

Ultra-Wideband VCO Tunes From 170 To 3900 MHz

THE MW500-1414 VCO TUNES from 170 to 3900 MHz while providing +11 dBm ±1.5 dB output. Tuning-voltage range is 0 to 18 V to cover this 2.3:1 bandwidth. This power output and bandwidth is provided by a compact 0.5 × 0.15-in. (1.27 × 0.381-cm) package, powered by +6 V at 35 mA. Spectral purity averages -107 dBc/Hz at 100-MHz offsets, with harmonics below -190 dBc. In addition, modulation bandwidth (tuning speed) is 12 MHz. This combination of features makes this VCO suitable for a wide range of test and measurement applications that require high-speed frequency agility and precision. Another application is the emerging interest in ultra-wideband (UWB) radio systems.

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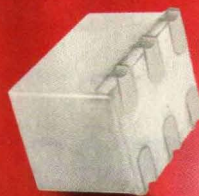
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RFID Emerges In Supply Chain

DESPITE THE SUCCESS of RF identification (RFID) technology within traditional application segments such as security/access

control, animal identification, automobile immobilization, and toll collection, the RFID industry has been highly focused

on finding RFID's next "killer application." The emergence of RFID in the supply chain over the last few years has led many to believe that supply chain may be the "killer application."

According to the recently published *Global Markets and Applications for RFID Equipment and Contactless Smart-card Systems*, 4th Edition by Venture Development Corp. (VDC), the global shipments of RFID hardware to support supply-chain management applications reached nearly \$89 million in 2002, with compounded annual growth estimated at slightly more than 38 percent through 2007.

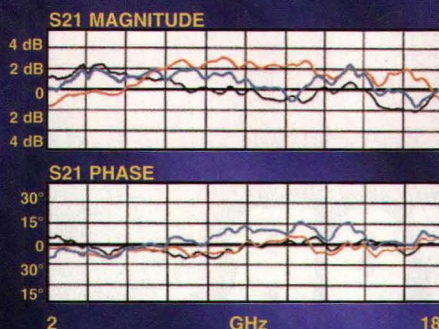
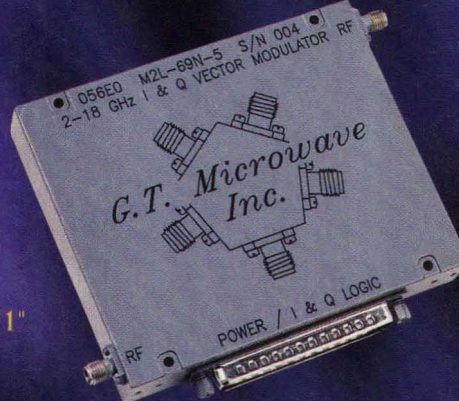
VDC research reveals that present RFID channels are largely underdeveloped and ill prepared to handle widespread adoption. For larger supply-chain management installations, qualified integrators are needed to take RFID system components and securely integrate them into a large-scale enterprise resource planning system or local databases. At present, the RFID market significantly lacks experienced, knowledgeable resellers and integrators. To expand their presence in the supply-chain market, many RFID vendors have been revamping their RFID education programs to include Web-based training, educational and training seminars, and sales coaching. However, further educational improvements are needed to capture opportunities within the supply chain and keep pace with the growing RFID market.

"Unfortunately, the revenue figures do not accurately reflect the level of interest and activity associated with RFID technology and the supply chain. While pilot activity and interest in supply-chain solutions increased steadily, few large-scale implementations were announced in 2002," states VDC project manager, Michael Liard. "The primary challenges facing RFID's expansion in the supply chain include: end-user price expectations, standards, and weak channel development." **MRF**

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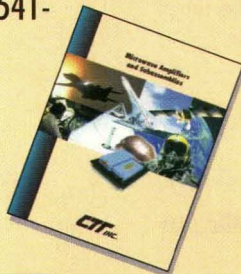
CTT's products contain the latest advancements in thin-film, GaAs FET MIC and MMIC technology. Having built thousands of products with power levels that range from a few milliwatts to hundreds of watts, CTT has become a leader in solid-state microwave amplification.

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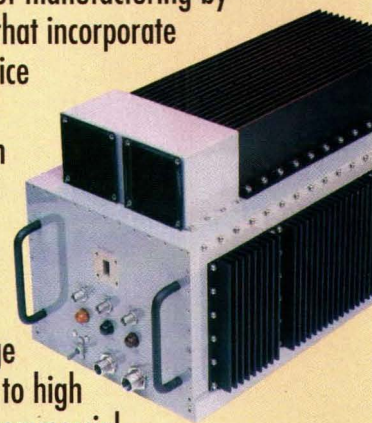
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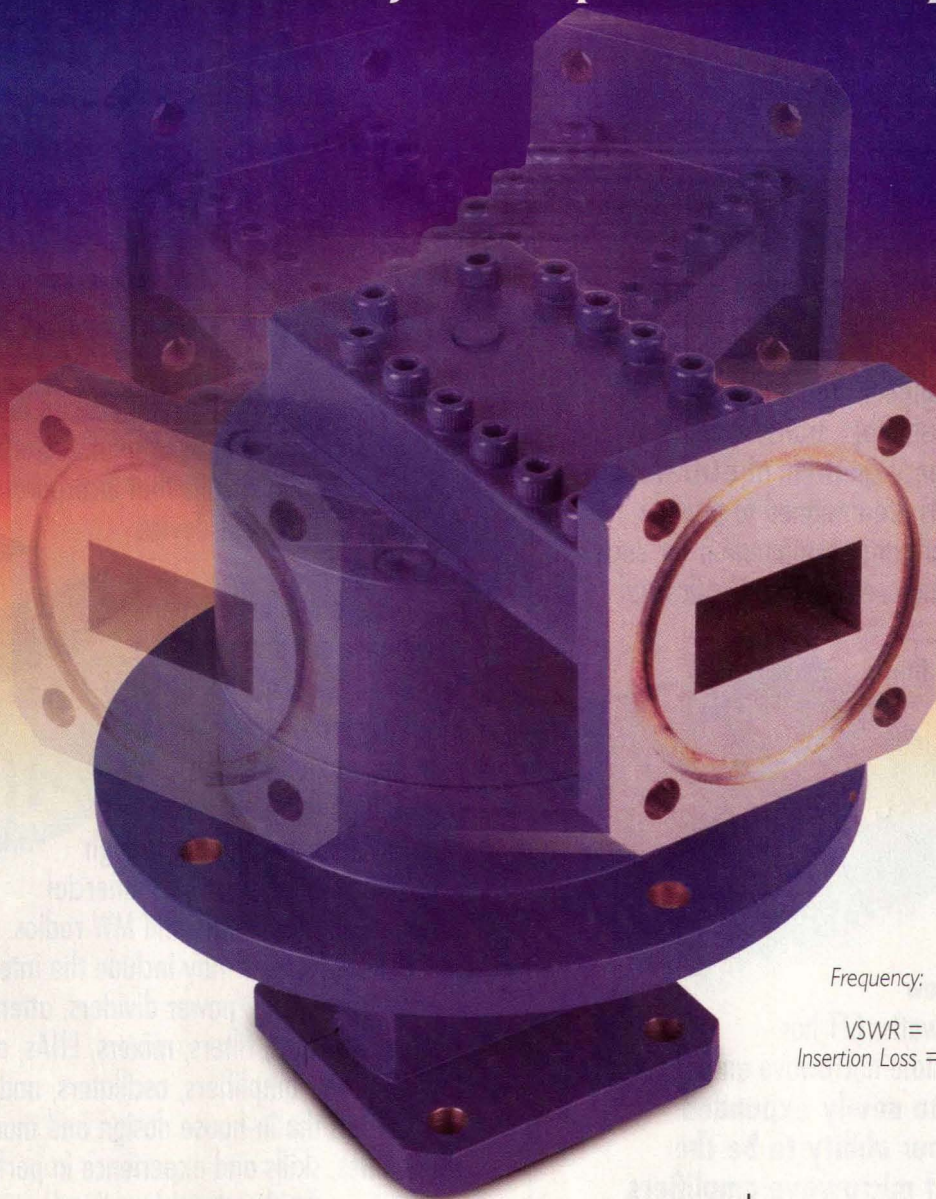
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AT&T—Was awarded a global networking service contract with Air China, the mainland's biggest airline. The multi-million-dollar deal includes both connectivity and network-management services covering 48 sites in Asia Pacific, Europe, the Middle East, Canada, and the US.

Air China asked AT&T to transform its corporate network to an IP-enabled global network platform, which will allow Air China to improve its operating efficiency and enhance internal communications. AT&T will provide global networking solutions that support Air China's international business operations, including its passenger-reservation system, customer management, frequent-flyer program, operation information, cargo, and airplane planning systems.

Under the agreement, AT&T will provide end-to-end global Frame Relay services supplemented by bilateral International Private Leased Circuits (IPLC) to link Air China's global network and data center in Beijing with its branch offices around the world. On the network-management side, AT&T will provide Managed Network Services from its Global Client Support Center (GCSC). As one of AT&T's regional hubs, the GCSC will constantly monitor the entire Air China network and provide early fault diagnosis and prevention services, as well as monitor the network performance to help ensure maximum efficiency.

CMC Electronics—Has received an order from Swiss International Air Lines to supply its high-gain Satellite Communications (Satcom) antenna system known as the CMA-2102 to Swiss International Air Lines for its fleet of 12 A340 aircraft. Deliveries will be completed by January 2004.

The CMA-2102 fully supports Swift64 high-speed data service without modification. It enables applications ranging from intranet Virtual Private Network (VPN) access for crews to multichannel voice services and fast e-mail for passengers. This new Inmarsat service has been designed to meet the needs of aircraft passengers, corporate users, and the flight deck, while making use of existing AeroH/H+ Satcom components already found on a large number of airline and corporate jet aircraft.

FRESH STARTS

Cougar Components Corp.—Has signed a Product Alliance Agreement with Marki Microwave, Inc., manufacturers of frequency mixers, converters, and doublers, DC to 40 GHz. Under the contract, Cougar adds to its current component capabilities of amplifiers, attenuators, limiters, lower-frequency mixers, and voltage-controlled oscillators (VCOs), by offering high-performance aerospace and defense customers a full line of microwave mixers. Cougar has developed sophisticated mixer test software to measure all mixer parameters

including rapid testing of harmonic intermodulation products as required for critical applications. Cougar, an ISO 9001 and MIL-PRF-38534 certified manufacturer, will manage the complete manufacture of Marki designs within ruggedized, hermetic-packaging configurations, including screening the mixers to MIL-M-28837, as required.

Tektronix—Opened a new manufacturing facility in Shanghai, China. The facility will be used for expansion in China, preparing for the future, as the country becomes a large segment of the test and measurement market. The building will initially serve as a primary manufacturing site for the region. Tektronix expects to continue to grow its presence over time by including even more services at the facility.

Micro Lambda Wireless, Inc.—Has moved to a larger facility. The new facility is approximately 20,000 sq. ft., an increase of 7000 sq. ft. from the old facility. The new address is: Micro Lambda Wireless, Inc., 46515 Landing Parkway, Fremont, CA 94538. Phone and fax numbers and the e-mail address remain the same: (510) 770-9221, FAX: (510) 770-9213, e-mail: sales@microlambdawireless.com.

Ansoft Corp.—Announced the release of a new device library based on NEC Compound Semiconductor Devices Ltd.'s Low Noise Bipolar Transistors for use within Ansoft Designer.™

The new device library contains electrical models of NEC Compound Semiconductor Devices Ltd.'s discrete low-noise bipolar transistor products for use in general-purpose high-frequency designs. These devices are key building blocks in the development of microwave frequency circuits and sub-systems found in many of today's commercial applications. This library improves design accuracy by supplying vendor-authorized device models directly as ready-to-use components.

This library is available from NEC Compound Semiconductor Devices Ltd.'s website at www.csd-nec.com and is easily downloaded and configured for use within Ansoft Designer. The website provides Ansoft Designer users with the ability to search for components by name, performance data, or other specification. NEC Compound Semiconductor Devices Ltd. will maintain model support and upgrades, while Ansoft will address all customer support for Ansoft Designer.

For further information, go to www.ansoft.com/products/hf/ansoft_designer.

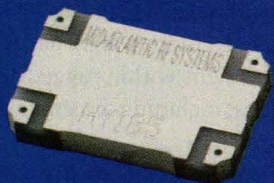
VXI Technology—Has relocated to a new location at 2031 Main St., Irvine, CA 92614. This 42,000-sq.-ft. facility is located near the Orange County John Wayne Airport.

The telephone and fax numbers and e-mail addresses will remain the same. Only the physical address has changed.

For more information on VXI Technology, see their website at www.vxitech.com.

Proxim Corp.—Announced that it is a principal member of the newly expanded WiMAX Forum and will also have a seat on the board of directors. WiMAX Forum is a non-profit corporation formed to promote and certify the compatibility and interoperability of devices using the IEEE 802.16 specifications. **MRF**

90° ±1° PHASE BALANCE

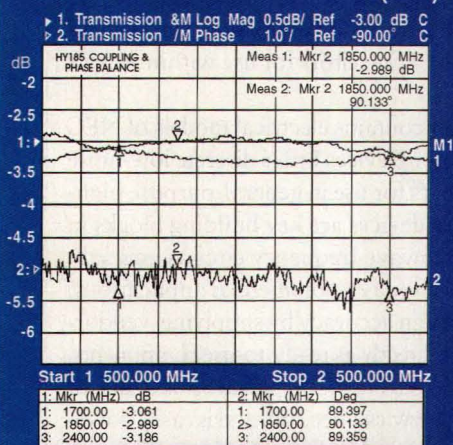


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DUPERRAY

ITT Industries, Cannon Promotes Duperray

ITT Industries, Cannon announced that OLIVIER DUPERRAY has been promoted to the role of senior vice president and COO-worldwide. Duperray had been serving as president of its global Switch Products business. He has been with ITT since 1983.

EMS Technologies, Inc.—BOBBY POTTS to director of operations for EMS Wireless; formerly director of quality for Scientific-Atlanta, Inc. Also, STEPHEN NEWELL to senior account manager for the SATCOM Aeronautical Group in Ottawa, Canada; formerly manager of avionics systems at AIRIA, Inc.

Enthone, Inc.—DR. YUN ZHANG to research director; formerly research director for tin and tin-alloy technologies at Technic, Inc. Also, AMY TSANG to director of marketing for Performance Coatings, Asia; formerly employed in various technical, manufacturing, and product-marketing management positions at Enthone.

Park Electrochemical Corp.—STEPHEN P. SCHAEFER to vice president of business development; formerly senior director of product technology.

Department of Energy—DR. LEONARD K. PETERS to director of the Department of Energy's Pacific Northwest National Laboratory; formerly vice provost for research at Virginia Polytechnic Institute and State University.

TRL Technology Ltd.—PETER MCKEE to managing director; formerly employed at Raytheon Systems Ltd.

QUALCOMM, Inc.—JING WANG to the position of chairman of QUALCOMM China; continues as senior vice president of QUALCOMM. Also, FRANK MENG to president of QUALCOMM China; formerly vice president of business development for QUALCOMM and head of QUALCOMM CDMA Technologies in China.

Elliptic Semiconductor—LARRY PERRON to the board of directors; remains as senior partner of Venture Coaches.

RLC Electronics, Inc.—TERRY OWENS

to sales manager; formerly vice president of sales with Microwave Filter Co.

Sarnoff Corp.—DR. HARPREET S. SAWHNEY to technical manager in Vision Technologies; formerly senior member of technical staff. Also, SRIDHAR KANAMALURU to technical manager for Microwave Systems; formerly technology leader for Broadband Wireless Networks. In addition, MICHAEL F. PATTI to technology director for the Integrated Circuit Systems Technology Center; formerly member of the technical staff.

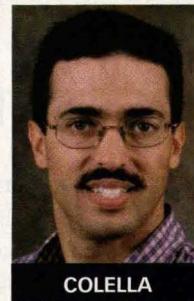
Sprint—GARY D. FORSEE to CEO and membership on the board of directors; formerly vice chairman of BellSouth Corp.

LXE, Inc.—PETER FAUSEL to vice president of marketing and business development; formerly president of Jacada, Inc., where he led the North American operations.

Eagleware Corp.—BILL CLAUSEN to product manager; formerly RF and microwave/optics applications engineer with HP/Agilent.



CLAUSEN



COLELLA

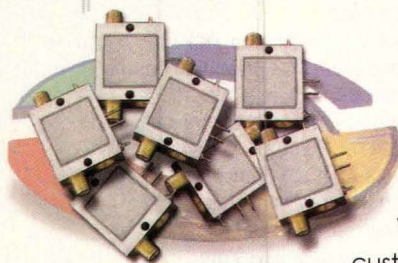
Spirex Corp.—MARK R. COLELLA to the position of vice president for sales and technology; formerly president of Forney, Inc. **MRF**

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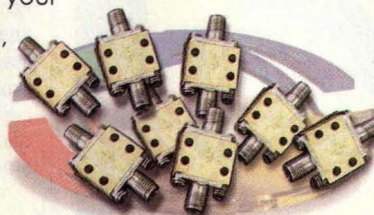
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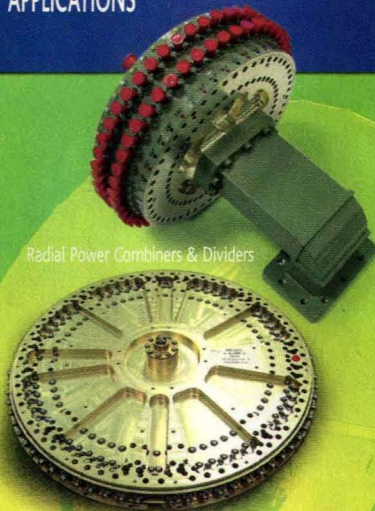
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CWC161-XXX*	16:1	1.4:1	0.30	0.25	0.50	1.5	3
CWC241-XXX*	24:1	1.4:1	0.30	0.25	0.85	2.0	3
CWC321-XXX*	32:1	1.4:1	0.50	0.30	0.85	2.0	4
CWC361-XXX*	36:1	1.4:1	0.50	0.30	0.95	2.0	4
CWC481-XXX*	48:1	1.4:1	0.60	0.40	0.95	4.0	5
CWC501-XXX*	50:1	1.4:1	0.60	0.40	0.95	4.0	5
CWC641-XXX*	64:1	1.4:1	0.60	0.50	1.20	5.0	8
CWC681-XXX*	68:1	1.4:1	0.60	0.50	1.20	5.0	8

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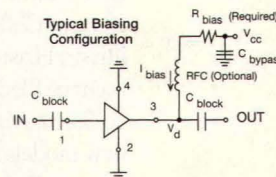
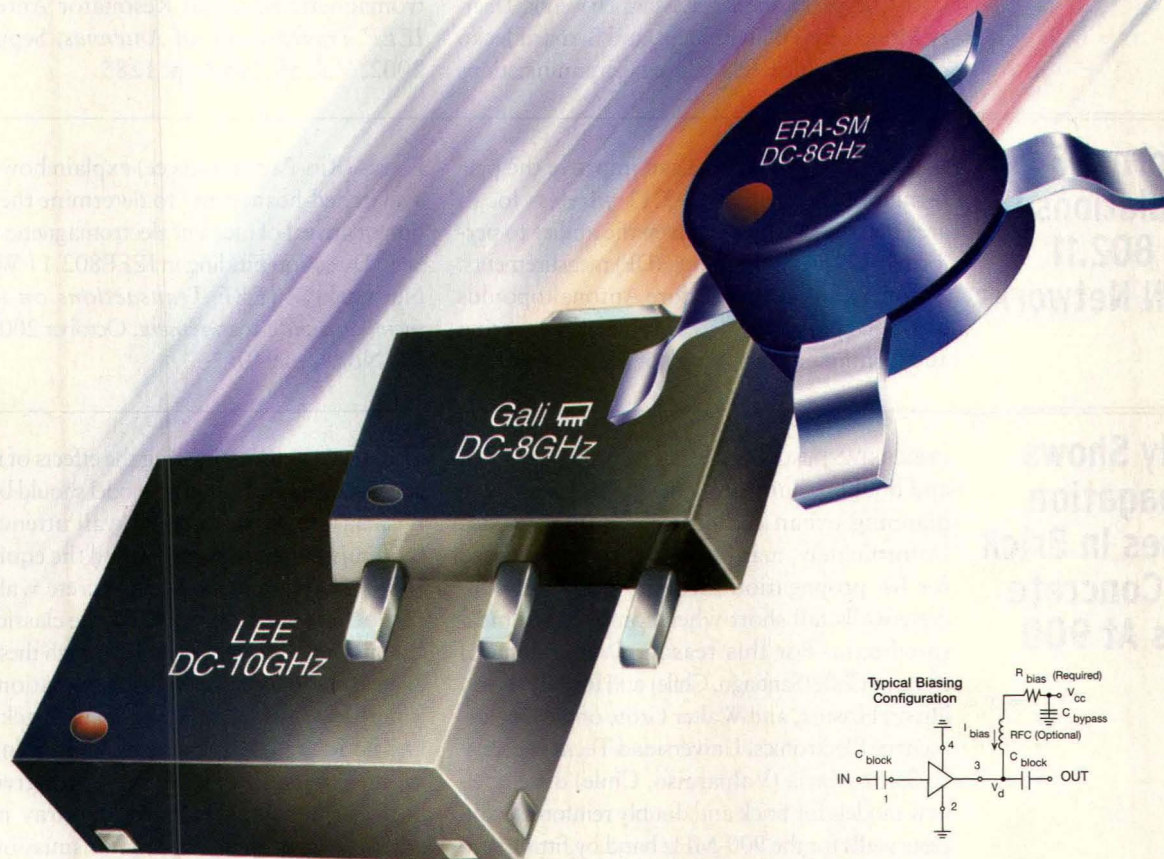
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376 Rev. A

EBG Materials Form Basis For New Resonator Antenna Design

ELECTROMAGNETIC BANDGAP MATERIALS (EBGs) have gained much attention among researchers for their promise in high-frequency circuits. These periodic structures are composed of metallic or dielectric elements that forbid the propagation of electromagnetic (EM) waves whose frequency falls within the frequency bandgap, although modes can be introduced by means of defects within the bandgap. Researchers Cyril Cheype, Cedric Serier, Marc Thevenot, Thierry Monediere, Alain Reineix, and Bernard Jecko of the Institut de Recherche en Communications

Optiques et Microondes (Cedex, France) used the properties of EBG materials to realize an EBG resonator antenna designed to work with about 3-percent bandwidth at 4.75 GHz. The device is fed by means of a patch antenna to excite desired modes. The antenna achieves better than 19-dB gain, with performance that is very competitive compared to classical antenna designs, including horns, reflectors, and lenses. See "An Electromagnetic Bandgap Resonator Antenna," *IEEE Transactions on Antennas*, September 2002, Vol. 50, No. 9, p. 1285.

Perform DF Calculations In IEEE 802.11 WLAN Networks

DIRECTIONAL ANTENNAS can improve the performance of an IEEE 802.11 wireless local-area network (WLAN), given the ability to perform basic direction-finding (DF) measurements. Antonis Kalis and Theodore Antonakopoulos of the Department of Electrical Engineering and Computer Technology of the University of

Patras (Rio-Patras, Greece) explain how to use a switched-beam array to determine the direction of arrival of incident electromagnetic waves. See "Direction Finding in IEEE802.11 Wireless Networks," *IEEE Transactions on Instrumentation and Measurement*, October 2002, Vol. 51, No. 5, p. 940.

Study Shows Propagation Losses In Brick And Concrete Walls At 900 MHz

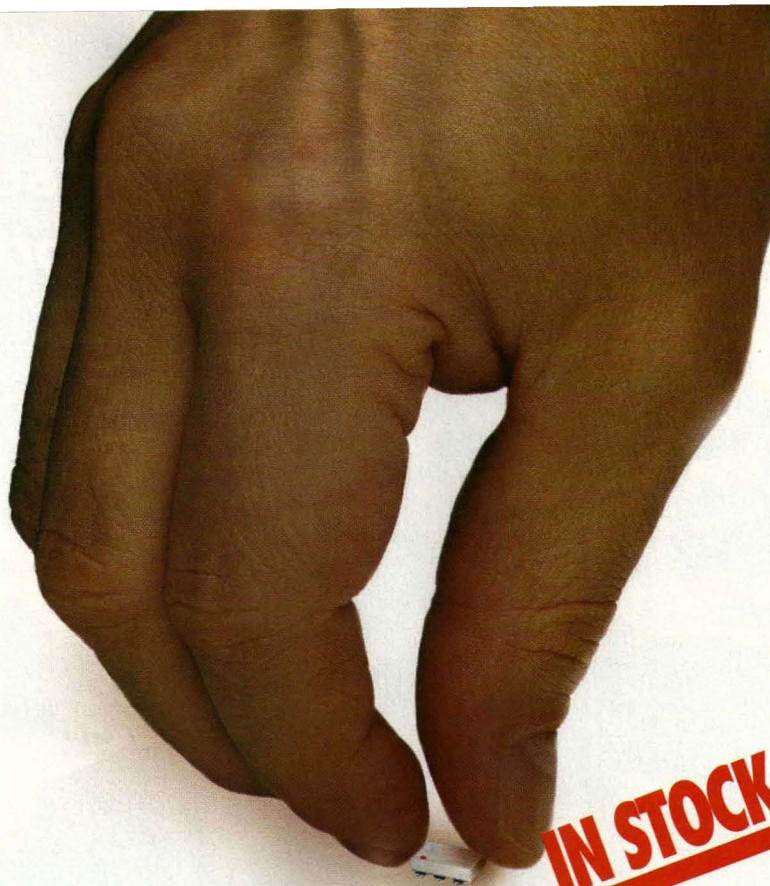
UNDERSTANDING RF PROPAGATION through brick and doubly reinforced concrete walls helps in planning urban cellular-telephone systems. Unfortunately, many of the models developed for RF propagation through brick and concrete walls fall short when compared to measured data. For this reason, Daniel Pena of Siemens Chile (Santiago, Chile) and Rodolfo Feick, Hristo Hristov, and Walter Grote of the Department of Electronics, Universidad Tecnica Fererico Santa Maria (Valparaiso, Chile) developed new models for brick and doubly reinforced concrete walls for the 900-MHz band by fitting simple ray-tracing models to empirical transmission data. They found that the single-ray model works well for lossy homogenous walls, in which the transmitted waves are quasiuniform. These same models are not as useful for nonuni-

form walls or for estimating the effects of reflected waves, where a multiray model should be used. By measuring the through-wall attenuation, the equivalent permittivity, and the equivalent conductivity of brick and concrete walls, the researchers were able to refine the classical single-ray model for propagation through these walls as well as for a one-path approximation using a multiray model. In particular, brick walls and concrete walls with two reinforcing steel meshes were studied with good agreement between the single-ray and multiray models and measurements for both transmissions and reflections. See "Measurement and Modeling of Propagation Losses in Brick and Concrete Walls for the 900-MHz Band," *IEEE Transactions on Antennas and Propagation*, January 2003, Vol. 51, No. 1, p. 31.

Meeting Modeling Challenges For Wireless Transceivers

DEVELOPING ACCURATE MODEL for wireless transceivers becomes more critical with time-to-market pressures and competitive wireless markets. For that reason, researchers Larry Dunleavy, Dann Benny Lassen, Terje Svensen, and Daniel Faria of the Center for Wireless and Microwave Information Systems, Department of Electrical Engineering, University of South Florida (Tampa, FL) examined the accuracy of a number of computer simulation programs when compared with measurements made on microwave vector

network analyzers (VNAs) and spectrum analyzers when considering 915-MHz transmitters (Tx) and receivers [Rx] (using software and test gear from Agilent Technologies). While good agreement was found between measurements and simulations for in-band performance, discrepancies were found for out-of-band response predictions. See "Computer-Aided Engineering Challenges for Wireless Transceivers," *IEEE RF and Microwave Computer-Aided Engineering*, January 2003, Vol. 13, No. 1, p. 86.



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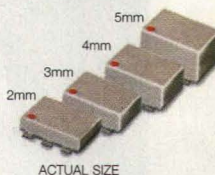
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ADE-3L	+3	0.2-400	5.3	47	10	4	4.25
ADEX-10L	+4	10-1000	7.2	60	16	3	2.95
ADE-1	+7	0.5-500	5.0	55	15	4	1.99▲
ADE-1ASK	+7	2-600	5.3	50	16	3	3.95
ADE-2	+7	5-1000	6.67	47	20	3	1.99▲
ADE-2ASK	+7	1-1000	5.4	45	12	3	4.25
ADE-6	+7	0.05-250	4.6	40	10	5	4.95
ADEX-10	+7	10-1000	6.8	60	16	3	2.95
ADE-12	+7	50-1000	7.0	35	17	2	2.95
ADE-4	+7	200-1000	6.8	53	15	3	4.25
ADE-14	+7	800-1000	7.4	32	17	2	3.25
ADE-901	+7	800-1000	5.9	32	13	3	2.95
ADE-5	+7	5-1500	6.6	40	15	3	3.45
ADE-5X	+7	5-1500	6.2	33	8	3	2.95
ADE-13	+7	50-1600	8.1	40	11	2	3.10
ADE-11X	+7	10-2000	7.1	36	9	3	1.99▲
ADE-20	+7	1500-2000	5.4	31	14	3	4.95
ADE-18	+7	1700-2500	4.9	27	10	3	3.45
ADE-3GL	+7	2100-2600	6.0	34	17	2	4.95
ADE-3G	+7	2300-2700	5.6	36	13	3	3.45
ADE-28	+7	1500-2800	4.5	30	8	3	5.95
ADE-30	+7	200-3000	4.5	35	14	3	6.95
ADE-32	+7	2500-3200	5.4	29	15	3	6.95
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ADE-1MH	+13	2-500	5.2	50	17	3	5.95
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ADE-10MH	+13	800-1000	7.0	34	26	4	6.95
ADE-12MH	+13	10-1200	6.3	45	22	3	6.45
ADE-25MH	+13	5-2500	6.9	34	18	3	6.95
ADE-35MH	+13	5-3500	6.9	33	18	3	9.95
ADE-42MH	+13	5-4200	7.5	29	17	3	14.95
ADE-1H	+17	0.5-500	5.3	52	23	4	4.95
ADE-1HW	+17	5-750	6.0	48	26	3	6.45
ADEX-10H	+17	10-1000	7.0	55	22	3	3.45
ADE-10H	+17	400-1000	7.0	39	30	3	7.95
ADE-12H	+17	500-1200	6.7	34	28	3	8.95
ADE-17H	+17	100-1700	7.2	36	25	3	8.95
ADE-20H	+17	1500-2000	5.2	29	24	3	8.95

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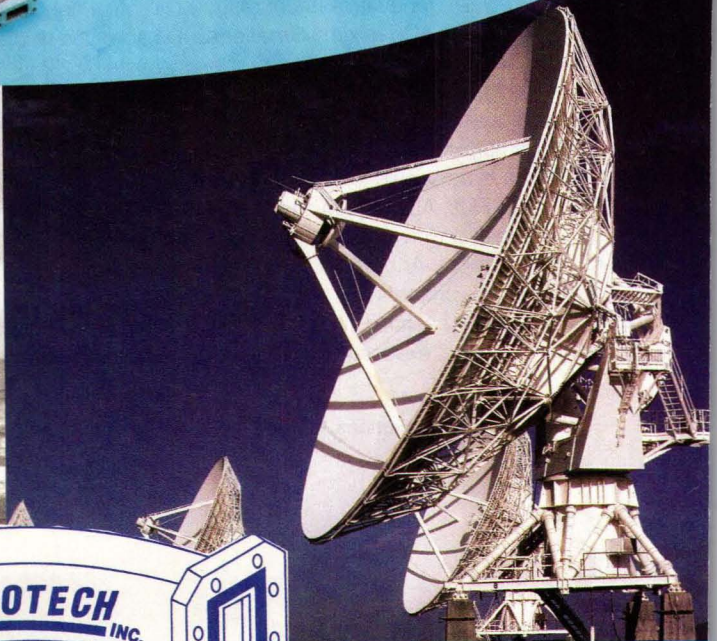
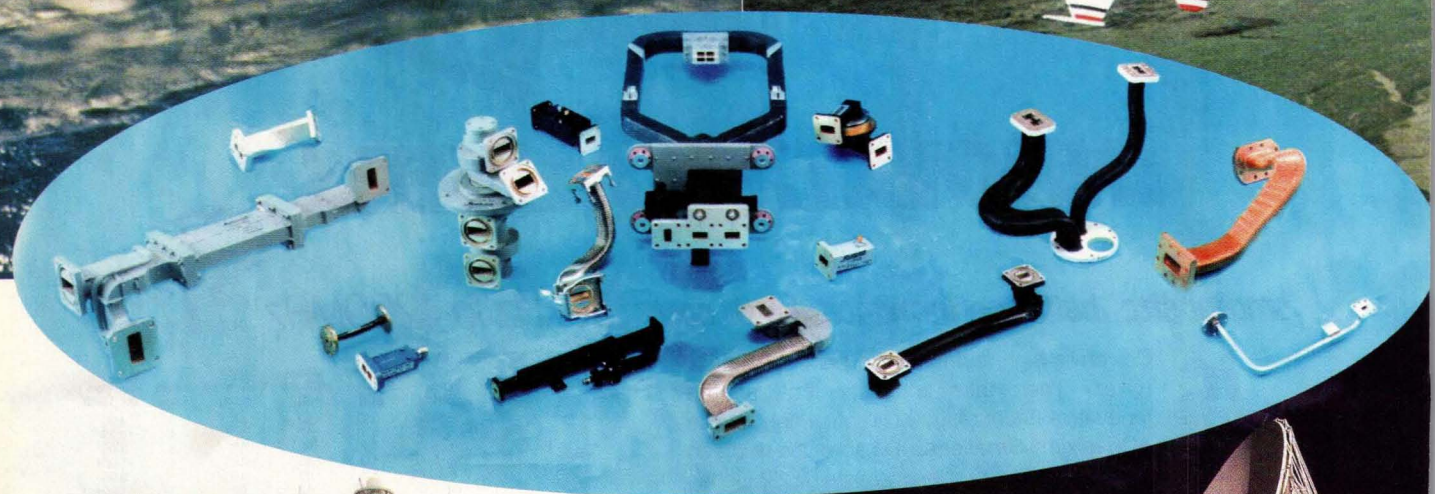
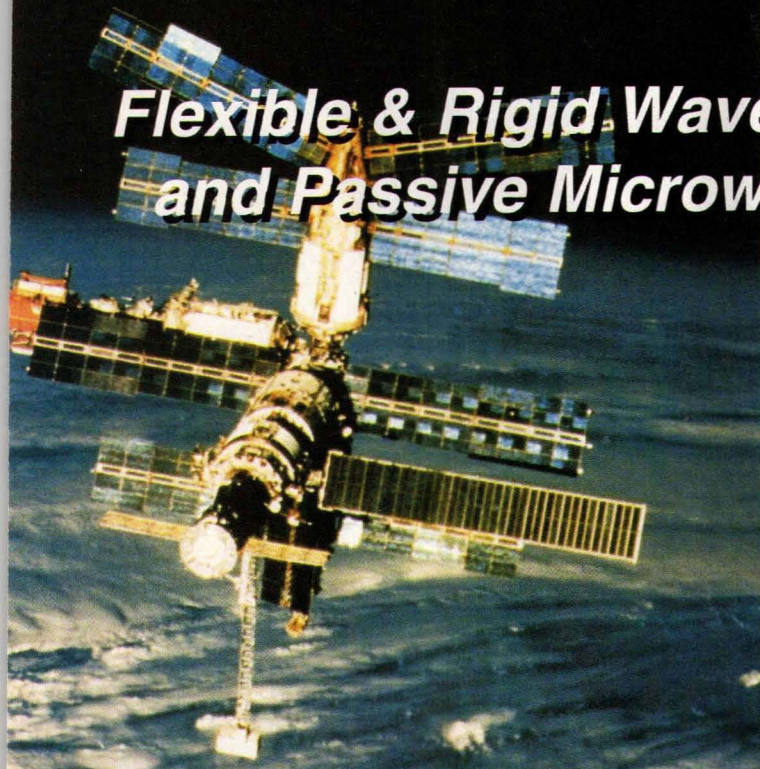


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Making Accurate Burst Measurements

Knowledge of pulse-timing and VNA measurement bandwidths can lead to accurate testing according to the burst-signal conditions of modern communications standards.

burst signals are now commonly used in commercial communications systems, such as Global System for Mobile Communications (GSM) cellular networks. Such signals are essentially pulsed RF waveforms characterized by long pulse widths and long periods (typically hundreds of microseconds to a few milliseconds long). Many components, particularly power amplifiers (PAs), must be tested

communications system will normally have a pulsed RF waveform associated with it. The waveform for a single GSM

channel (e.g., [1]) is shown in Fig. 1 and some other systems have parameters within an order of magnitude of those shown. Relative to pulsed radar systems, a measurement challenge from earlier years, the pulse widths for GSM and similar systems are large, duty cycles are fairly large and periods are long. Fortunately, a vector network analyzer (VNA) can be a useful tool for rapidly extracting this information since it has fast-moving receivers (Rx) and, assuming it has an appropriate per point analog triggering function, can make all of the measurements listed above among others. Some guidelines for performing burst measurements with a VNA follow.

A time-division-multiplexed com-

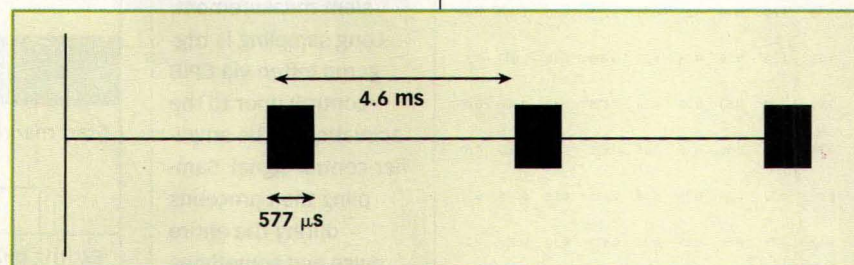
munications system will normally have a pulsed RF waveform associated with it. The waveform for a single GSM channel (e.g., [1]) is shown in Fig. 1 and some other systems have parameters within an order of magnitude of those shown. Relative to pulsed radar systems, a measurement challenge from earlier years, the pulse widths for GSM and similar systems are large, duty cycles are fairly large and periods are long.

Many PAs used in such communications systems are heavily optimized for efficiency and spectral purity for a given output power in this pulsed format. As such, many do not operate correctly, if at all, with a CW waveform that would typically be used for a measurement. In addition, the amplifier (or

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1. This illustration shows the RF envelope of a single GSM channel. The duty cycle is about 12.5 percent.



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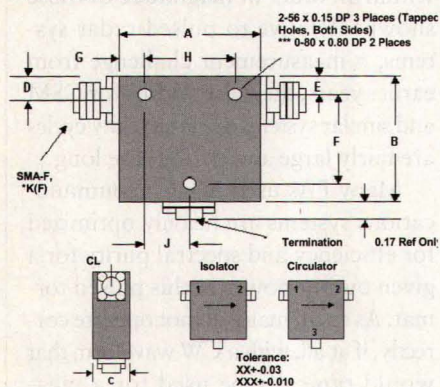
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D3I0120	1.7-2.0	20	.40	1.25	3	\$210.00
D3I0223	2.0-2.3	20	.40	1.25	3	\$210.00
D3I0240	2.0-4.0	18	.50	1.30	1	\$215.00
D3I0260	2.0-6.0	14	.80	1.50	1	\$250.00
D3I0280	2.0-8.0	10	1.50	2.00	1	\$395.00
D3I0360	3.0-6.0	19	.40	1.30	2	\$195.00
D3I0480	4.0-8.0	20	.40	1.25	3	\$185.00
D3I0612	6.0-12.4	17	.60	1.35	6	\$195.00
DM6018	6.0-18.0	14	1.00	1.50	11	\$275.00
D3I0711	7.0-11.0	20	.40	1.25	4	\$185.00
D3I0712	7.0-12.0	20	.40	1.25	4	\$205.00
D3I0718	7.0-18.0	15	1.00	1.50	5	\$225.00
D3I0812	8.0-12.4	20	.40	1.25	4	\$180.00
D3I0816	8.0-16.0	17	.60	1.35	5	\$205.00
D3I0820	8.0-20.0	15	1.00	1.45	5	\$230.00
D3I1020	10.0-20.0	16	.70	1.40	5	\$220.00
D3I1218	12.0-18.0	20	.50	1.25	5	\$180.00
D3I1826	18.0-26.5	18	.80	1.40	5	\$225.00
D3I1840	18.0-40.0	10	2.00	2.00	5*	\$1300.00
D3I2004	20.0-40.0	12	1.50	1.65	5*	\$950.00
D3I2640	26.5-40.0	14	1.00	1.50	5*	\$700.00

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D3C0116	1.4-1.6	20	.40	1.25	8	\$235.00
D3C0118	1.6-1.8	20	.40	1.25	3	\$210.00
D3C0120	1.7-2.0	20	.40	1.25	3	\$210.00
D3C0223	2.0-2.3	20	.40	1.25	3	\$210.00
D3C0240	2.0-4.0	18	.50	1.30	1	\$215.00
D3C0260	2.0-6.0	14	.80	1.50	1	\$250.00
D3C0280	2.0-8.0	10	1.50	2.00	1	\$395.00
D3C0360	3.0-6.0	19	.40	1.30	2	\$195.00
D3C0480	4.0-8.0	20	.40	1.25	3	\$185.00
D3C0612	6.0-12.4	17	.60	1.35	6	\$195.00
DMC6018	6.0-18.0	14	1.00	1.50	11	\$275.00
D3C7011	7.0-11.0	20	.40	1.25	4	\$185.00
D3C7018	7.0-18.0	15	1.00	1.50	5	\$225.00
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D3C8020	8.0-20.0	15	1.00	1.45	5	\$230.00
D3C1218	12.0-18.0	20	.50	1.25	5	\$180.00
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other DUT) may exhibit unusual behavior at the start of the pulse or some other time subset. It may also be of interest to measure the DUT during a certain subinterval of the pulse (termed 'profiling').

Some PAs, particularly in handsets, can operate CW but they may be put in an inactive state (via a control voltage) during the off-periods of the cycle. How the amplifier responds, in a transient sense, at the beginning and end of the pulse period (being turned active and inactive) are very important and represent another example of profiling.

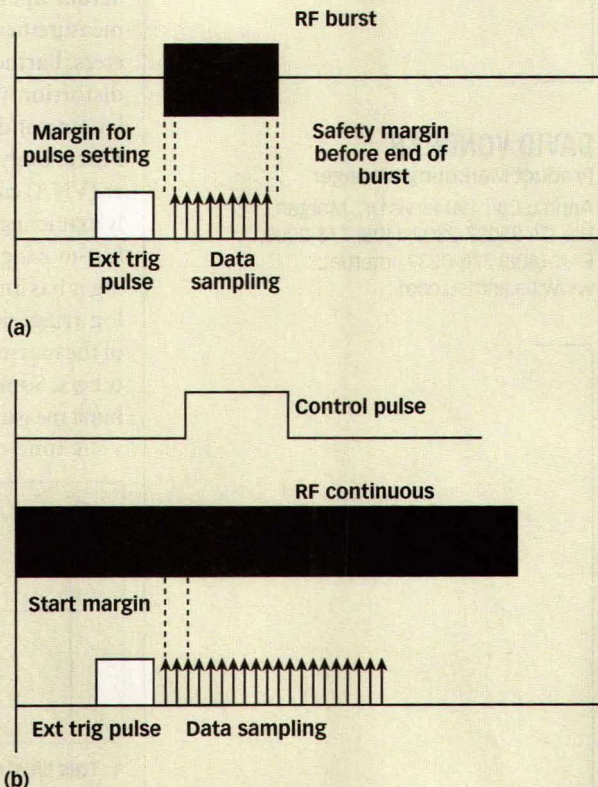
S-parameter and harmonic measurements are of somewhat obvious interest. Power sweeps of gain and output power, subsets of the S-parameter measurement class, are often of greatest interest for PAs. In terms of executing the burst measurement, however, there is little difference from the general case. Relative to the control-voltage profiling example, the power and frequency are usually constant; it is a pure transient response of the S-parameter or other quantity that is of interest.

Among other PA measurements is intermodulation distortion [IMD] (e.g., refs. 2 and 3) although its related cousin, adjacent-channel power rejection (ACPR) (e.g., refs. 4 and 5) is increasingly important. Both of these linearity measurements have the characteristic of some strong applied signal with a resulting distortion signal at a different, yet close, frequency. For a highly linear DUT, this is then a high-dynamic-range measurement with spot power levels below -70 dBc sometimes of interest.

One approach to doing this measurement would be with a spectrum analyzer and just consider the main lobe of the response generated by the pulse train. For IMD measurements with small separations, this may become impractical. There are also limitations in flexibility and amplitude accuracy, even with triggered measurements (similar to that discussed in the next section), with such a setup. For S-parameter measurements, the spectrum-analyzer approach cannot offer error correction, which further impairs accuracy. Previous techniques involving CW filtering of the main lobe of the pulse

2. The top illustration

(a) shows direct data sampling during an RF pulse. Some pulse-generator settling time is required (hopefully less than the DUT), and sampling can usually continue within a few microseconds of the end of the pulse. The bottom illustration (b) shows direct data sampling during a CW transient measurement. Long sampling is triggered (often via GPIB control) prior to the activation of the amplifier control signal. Sampling then proceeds during the entire pulse and sometimes afterward.





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MCA1-42LH	10	1000-4200	6.0	38	7.45
MCA1-60LH	10	1700-6000	6.3	30	8.45
MCA1-24MH	13	300-2400	6.1	40	6.95
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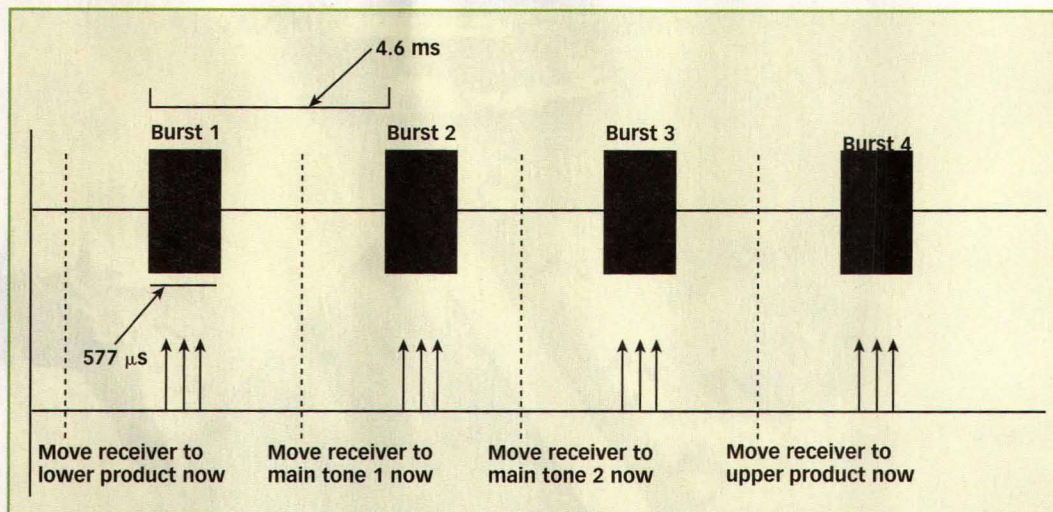
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spectrum⁶ appear impractical due to the duty cycles and pulse widths involved in most relevant communications systems. This leaves direct data sampling during the pulse-on period as a potentially accurate and flexible approach.

In the case of examining transient amplifier behavior during a control pulse, direct data sampling is perhaps the only choice. Any kind of Rx can potentially make this measurement if it is scalar (e.g., gain) and the Rx can take samples fast enough. If error correction is required (S-parameters) or IMD measurements



3. This is the measurement scheme for a full IMD sequence under GSM conditions. One of the four tones is measured during each burst to allow for more flexibility in how many samples are acquired per tone and hence better dynamic-range possibilities.

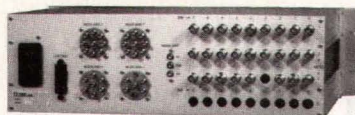
are needed, then a network-analyzer approach with direct data sampling is a valid option.

The measurement of a single signal

using the direct-data-sampling technique is illustrated in Fig. 2 for both the (a) RF burst scenario and the (b) CW transient measurement. A trigger pulse

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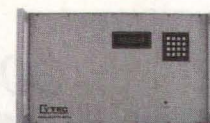
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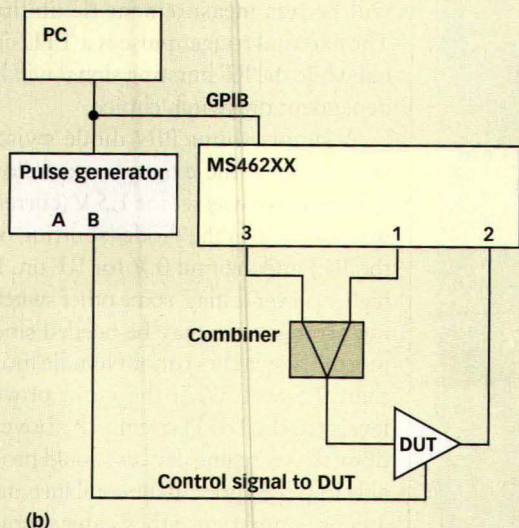
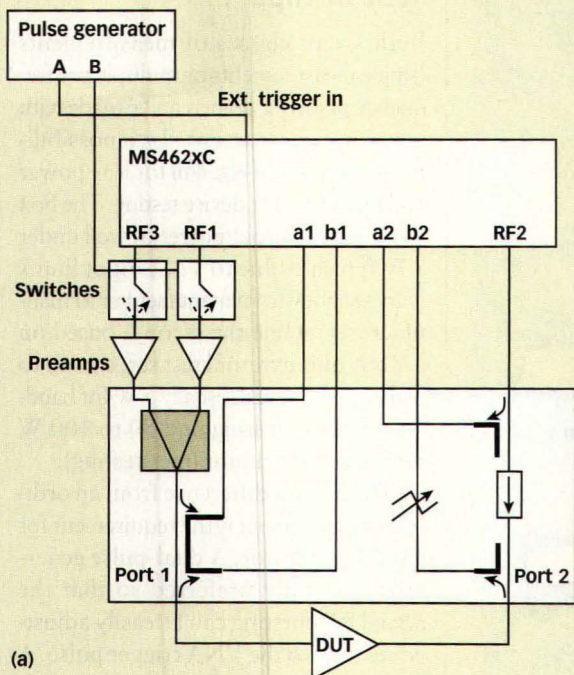
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is used to trigger the measurement and it must be synchronized with the RF burst (often by using a dual-pulse generator). The trigger pulse orders the VNA or vector-network measurement system [VNMS] (assuming the instrument has been set to accept external triggering) to perform the measurement and the instrument's analog-to-digital converters (ADCs) begin converting after a certain latency delay. The pulses must be arranged so that the burst is settled

by the time the latency ends. For the setups to be discussed here, the pulse settling is very short ($< 1 \mu\text{s}$) but the DUT behavior early in the pulse may be of interest. The sampling must not continue on too long or it will overrun the burst and the resulting composite data will probably be corrupt. Since the measurement time is constrained by the GSM pulse definition (assuming no more than one burst is needed for a single tone measurement), reducing the

averaging will not speed overall measurement throughput.

Figure 2 shows the measurement of one tone (as would be appropriate for an S-parameter or harmonic measurement) while a complete IMD measurement typically requires the measurement of four tones (one for each of the two main tones, one for the lower product and one for the upper product). While conceivably this could be done within one burst, there may not be enough time to get enough samples for the dynamic range required of the measurement.

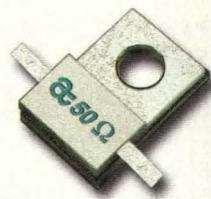


4. Two example setups are shown for swept measurements (a): for lower-power applications (with input power of less than +20 dBm to the DUT, or less than +5 dBm for IMD measurements, and less than +30 dBm output power), and for higher-power testing using an external power-amplifier test set. Another example setup is shown (b) for transient measurement of a DUT under CW conditions.

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Most modern Rx's use some form of digital intermediate-frequency (IF) filtering so narrower IF bandwidths (and hence more dynamic range) require more samples. Thus the measurement configuration of **Fig. 3** will be used where one tone is measured per burst.

The picture shown would then repeat for the next frequency or power point (by this we mean both main tones move to new frequencies or new powers assuming both tones are sweeping³).

For a control-response-transient measurement (CW and constant power,

a control voltage to the DUT is synchronized with the pulse-timing pattern), it is simpler to just continuously sample data. Each set of N samples (dependent on IFBW) represents one time point allowing a time-trace to be taken. This allows maximum time resolution of the DUT's response. In the case of IMD measurements, it is perhaps most accurate to measure one tone on each subsequent control pulse.

Test Setups

Both swept classes of measurements (single measurements or multiple per frequency or power point) can be made with a variety of test setups. Two possibilities are shown in **Fig. 4(a)** for low power and higher power device testing. The first is limited to output power of well under 1 W typically due to VNA input limits (varies slightly from manufacturer to manufacturer) while the second, based on a VNA plus external test set, can go to higher power levels (usually 5 W for handset-amplifier testing and 50 to 100 W for base-station-amplifier testing).

The biggest difference from an ordinary measurement is the requirement for a pulse generator. A dual-pulse generator is usually preferred so that the actual RF bursting can be easily adjusted relative to the VNA trigger pulse. A fixed-delay generator to create the trigger pulse is also a possibility but there will be less measurement flexibility. The external trigger pulse is a TTL signal while the RF bursting signal will be dependent on switch choice.

A simple shunt PIN diode switch was used for the examples to follow and the drive was set for 1.5 V (current limited) to bias the diode (shutting off the RF) and around 0 V for RF on. In higher power testing, some other switching arrangement may be needed since most PIN switches cannot handle more than 0.5 to 1 W. If the input power levels (to the DUT) rise much above 0 dBm, the switching devices should probably be placed prior to the combiner (and any pre-amps) since they can generate their own intermodulation products. Generally the reverse signals are not

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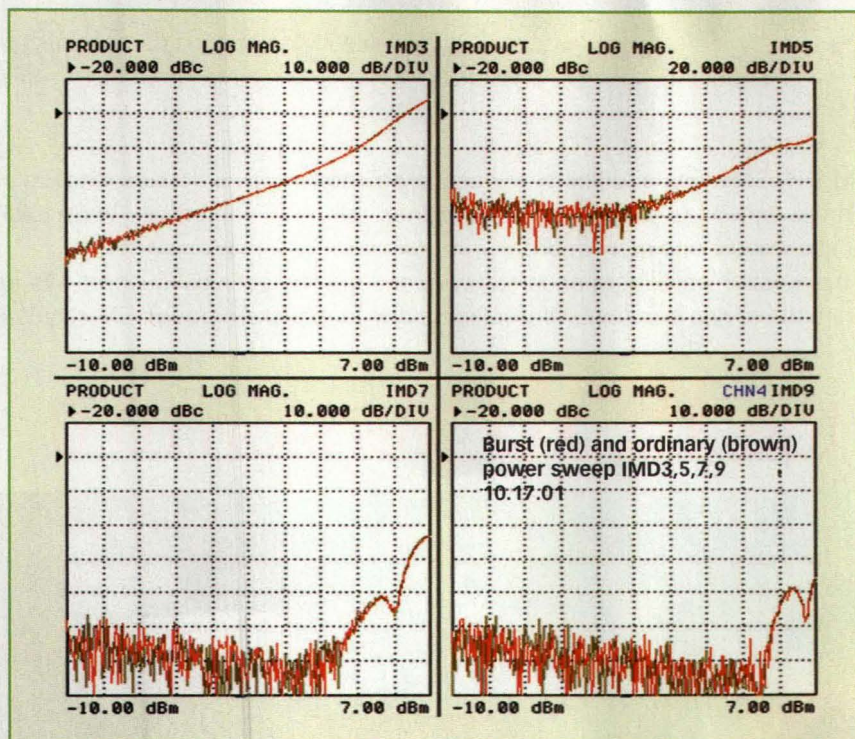
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5. This power-sweep example shows a well-behaved amplifier with and without GSM bursting. The third-, fifth-, seventh-, and ninth-order IMD products for this amplifier are not a function of burst versus CW signal.

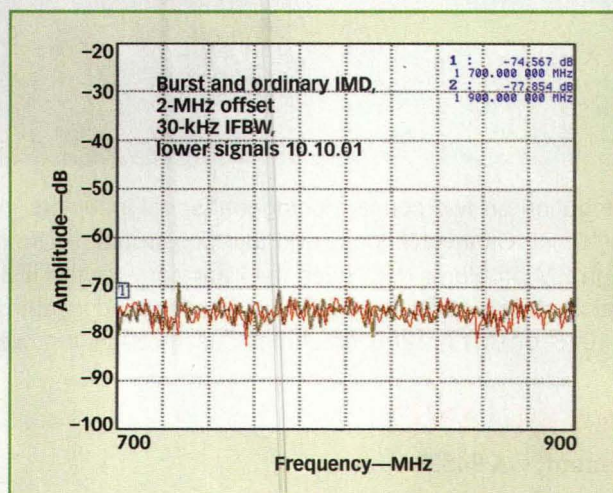
burst since most DUTs would not be dependent on the waveform of a small reverse signal. This path could also be switched, however, if desired.

An example setup for transient measurements under RF CW conditions is shown in Fig. 4(b). In this case, an RF switching mechanism is obviously not needed but the pulse generator is generally still required to create the control

pulse for the DUT. In many cases, the operating mode of the VNA may need to change to accommodate the fast sampling required for this measurement so GPIB may be used to trigger this mode and the sweep and to synchronize the control pulse with the sweep.

With the base setup in place, there are a number of measurement details to consider. The burst operation itself

rarely limits bandwidth since very broadband switch assemblies (greater than 10-GHz bandwidth) are available. In the case of the IMD-compatible setups shown in Fig. 4, the combiner (often a Wilkinson structure) is usually the limiting factor for lower power setups. These are now available covering a few hundred megahertz to at least 5 GHz so reasonable bandwidth measurements

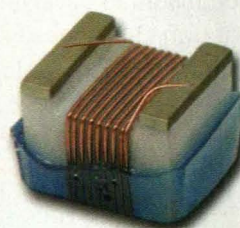


6. This swept-frequency example shows third-order intermodulation performance with and without GSM bursting.

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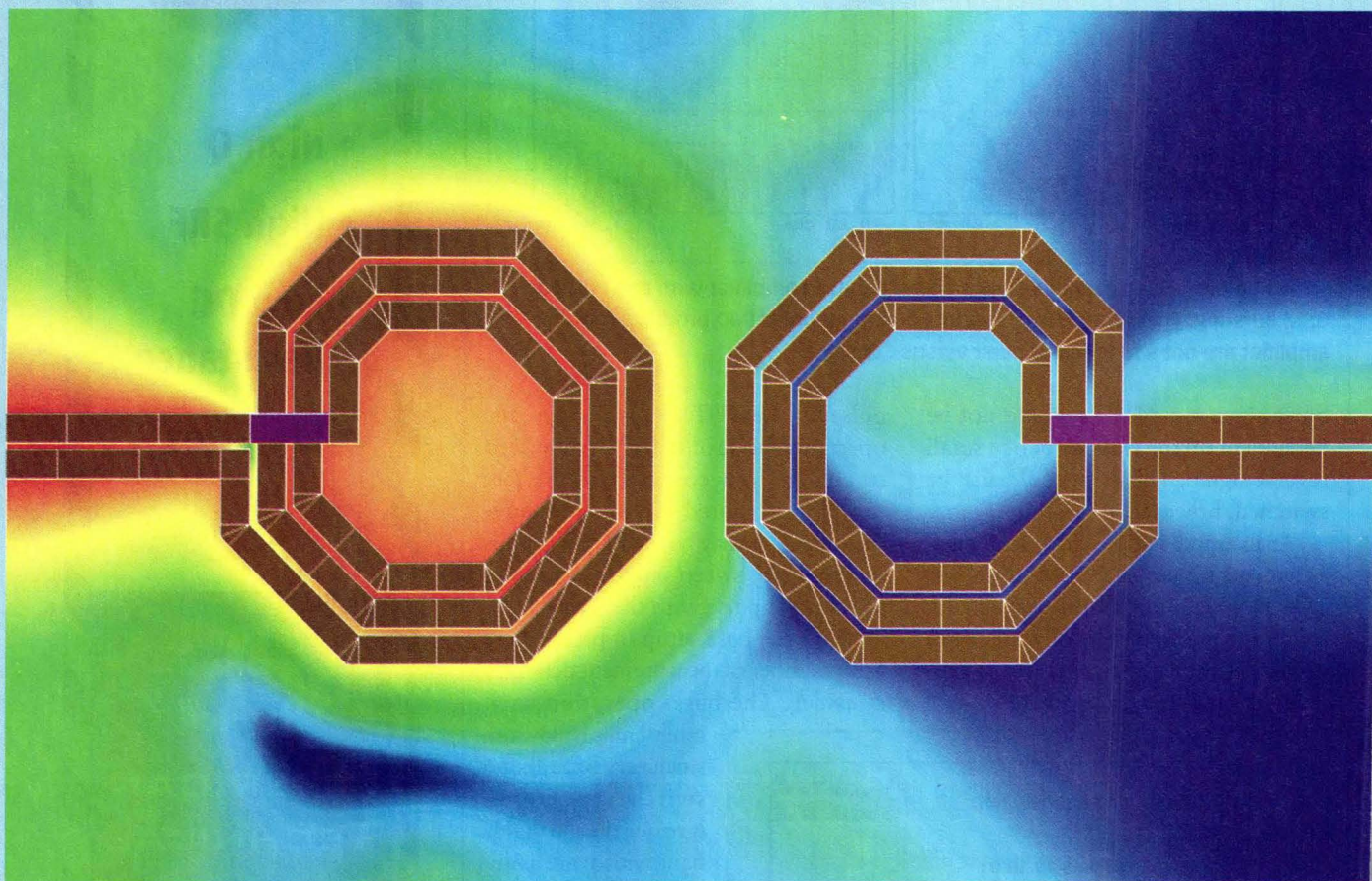
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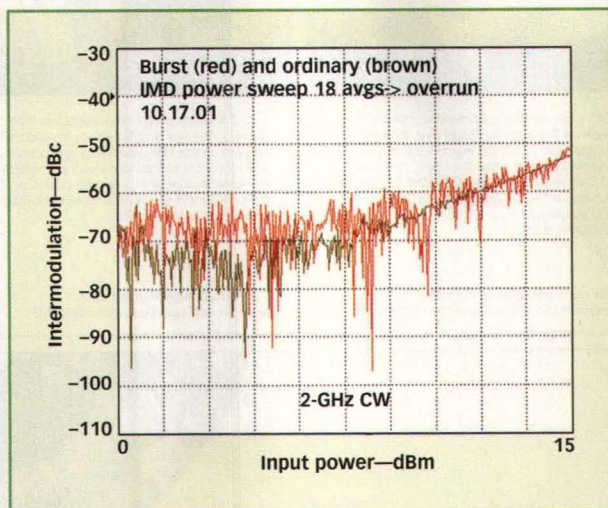
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7. This plot shows a power-sweep measurement of an amplifier when too many measurement samples are attempted.

are possible. In higher-power setups (for base-station devices for example), the passive components and the DUT itself will often be narrowband for power-handling reasons.

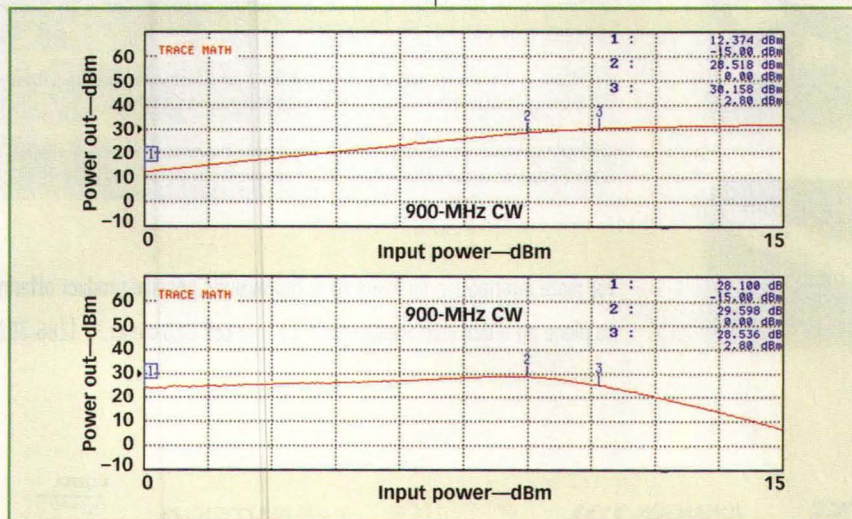
The test setups were designed to allow some flexibility in timing between the RF pulse and the measurement trigger pulse. There is some latency between when the trigger pulse occurs and when the measurement starts in the VNA (50 to 100 μ s in the MS462XX family of VNAs) so there is no need for the RF pulse to start until about that delay after the trigger. A small amount of time (less than 10 μ s usually) might be needed for

the switch to settle. If DUT bias switching is being used, the time constants of that bias network are relevant.

The length and period of the RF pulse are set by the standard in question (GSM, for example) so those parameters are not free. The measurement time (i.e., number of samples) can be changed as long as the measurement is within the RF pulse. In the MS462XX, the maximum number of averages is about 17 in a

30-kHz bandwidth for a clean measurement on a GSM pulse. A smaller number can be used to profile the behavior of the DUT in a certain part of the pulse (as will be shown in an example).

Power levels are usually set by the DUT requirements. As discussed previously, the only new requirements are to keep the drive levels at the switch low enough to avoid overdriving them. In the case of transient measurements, the detailed shape of the control pulse (ramp speeds) are often of critical interest and may be specified by appropriate communications standards. This is usually of less interest in the case of swept (RF burst)



8. This plot shows a swept-power S-parameter measurement of a power amplifier under burst conditions (amplifier-gain expansion is obvious).

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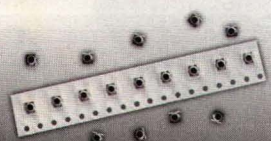
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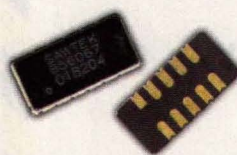
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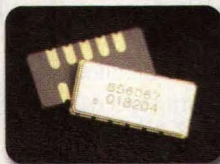


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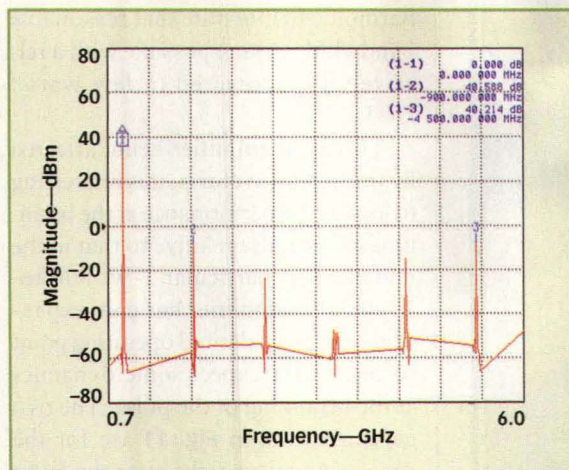
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9. This plot shows a burst harmonic measurement of a power amplifier (from Fig. 8). The input is fixed at 900 MHz while the VNA receiver is swept over the fundamental plus second through sixth harmonics.

measurements.

To gain confidence in the method, measurements on well-behaved amplifiers (wideband, shallow compression, no gain expansion, and no linearizers) and passive devices will first be presented with and without the switch present. Since the IMD level of such a device should not be dependent on the pulse form or timing, the measured results with and without switching should be the same. Examples confirming this are shown in Figs. 5 and 6.

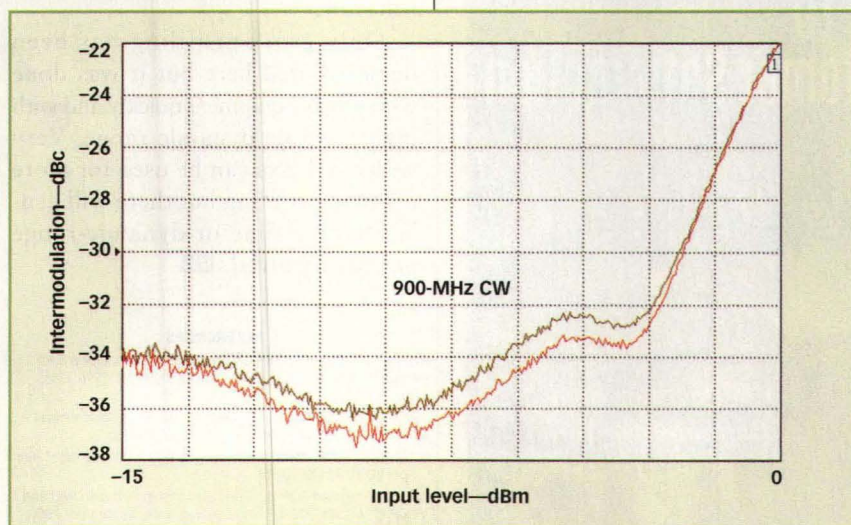
Figure 6 also illustrates the avail-

able dynamic range with such a burst measurement. While these graphs suggest that the timing is being done correctly, it may be illustrative to view an example where it is not. Figure 7 shows a linear-amplifier power-sweep measurement where the measurement is continued beyond the end of the burst. The resulting noisy data indicates the measurement failure.

A more conventional S-parameter measurement of a 900-MHz PA was performed next. This will be a power-sweep measurement and the output power and gain are shown in Fig. 8. The gain expansion in this particular amplifier is obvious. Since the gain is a strong function of RF drive in this region, it follows that this amplifier's performance will be a function of the burst waveform and hence it is important that it be tested with the appropriate input signal.

Harmonic measurements of an amplifier under these measurement conditions may also be of interest. A single-frequency, single-power measurement is shown in Fig. 9 covering up to the sixth

Figure 9 also illustrates the avail-



10. This plot shows a burst swept-power third-order-intermodulation measurement of a handset power amplifier. The upper trace corresponds to a measurement at the beginning of the burst (after the pulse itself has settled) while the lower trace corresponds to a measurement in the middle of the burst.

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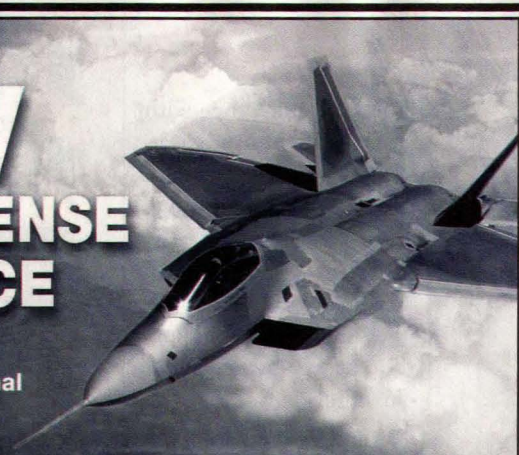
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DESIGN

harmonic to illustrate that reasonable bandwidth is quite possible with a relatively low-powered (a few watts) DUT.

To see an amplifier being affected by the burst waveform, it is interesting to look at the performance at the beginning of the pulse relative to that in the middle. This particular 1-W, 800-to-1000-MHz amplifier has gain expansion prior to its desired operating point so one might expect some dynamics at the beginning of the pulse. The two measurements in **Fig. 11** are for the data being taken right after the burst starts and then midway into the burst (300 μ s). The 1-dB peak difference is interesting in that it is somewhat averaged (measurement over the first 65 μ s of the pulse) so it under-represents the peak. It also suggests some potential spectral quality problems for the first portion of the burst cycle for certain power levels.

At very low power levels and closer to compression (the right and left extremes of Fig. 10), one sees a much-smaller difference, which might be consistent with DUT power being a less violent function of input power in these ranges. It is important to note that the measurement early in the burst was done after the input burst itself had settled so the difference is due to the DUT behavior.

Only gross profiling has been demonstrated here but it was done with simple equipment quickly and with high possible dynamic range. Very wideband Rx's can be used for more detailed profiling but there will generally be a time or dynamic-range penalty involved. **MRF**

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6. Model 360B Pulse System Operation Manual, Anritsu Co., Morgan Hill, CA.
7. Steve Cripps, *RF Power Amplifiers for Wireless Communications*, Artech House, Norwood, MA, 1999, Chap. 10.

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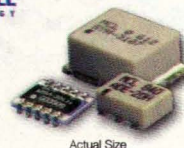
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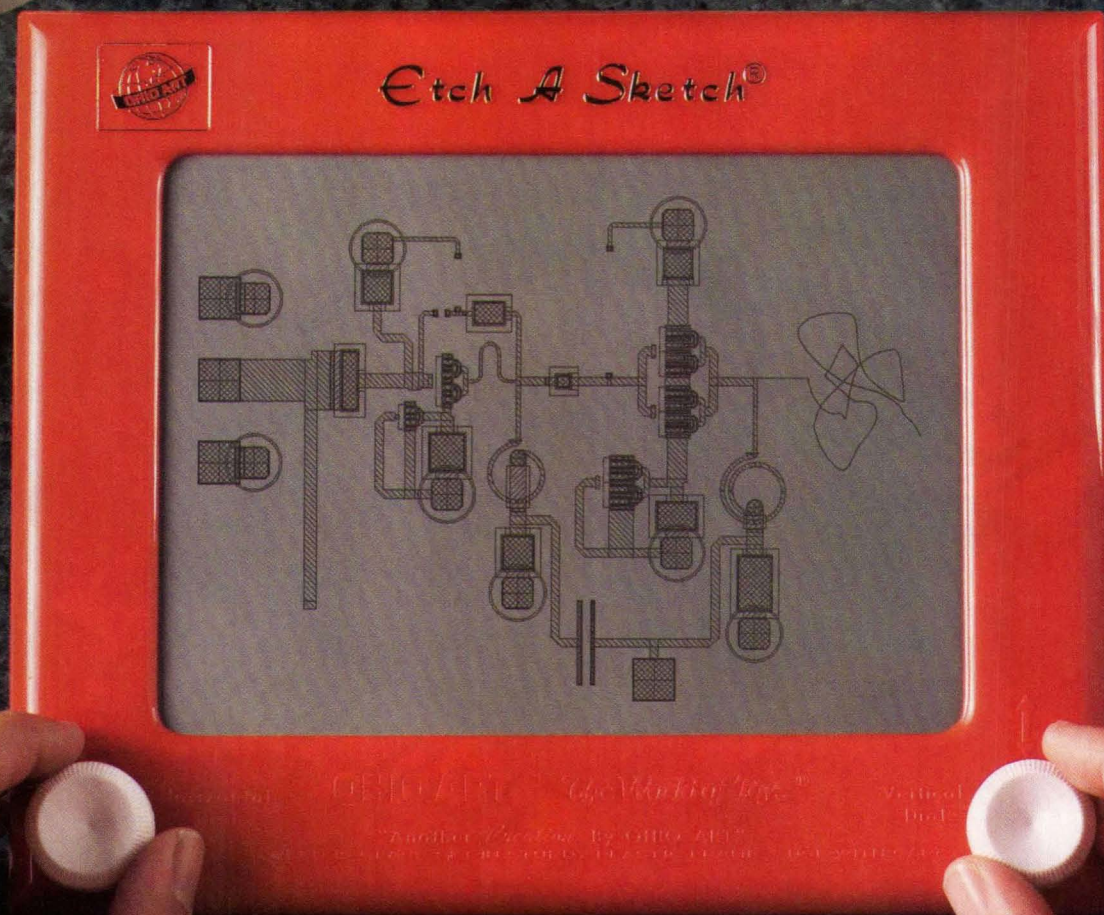
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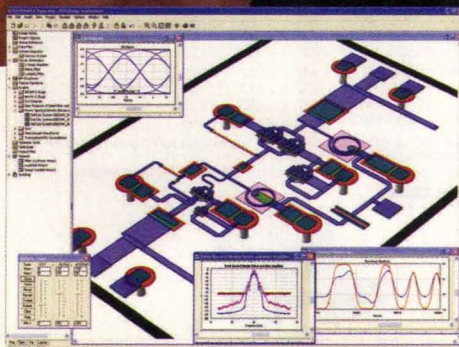
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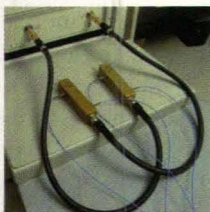
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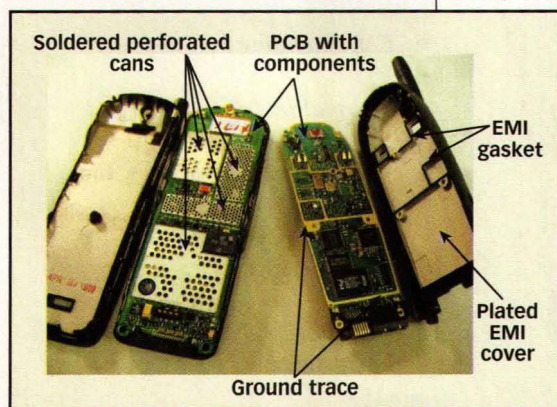
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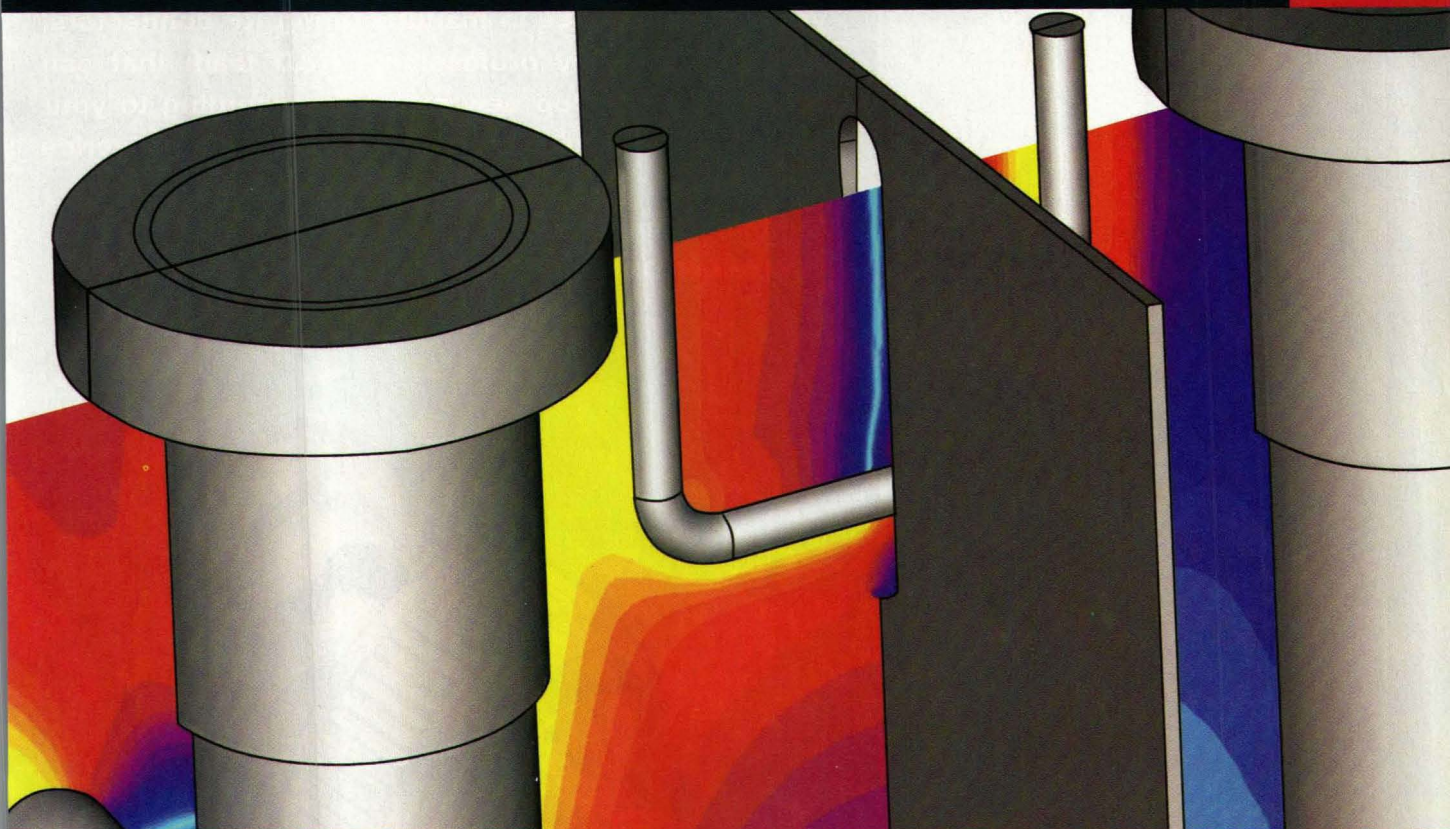


1. Traditional PCB shielding methods include soldered perforated cans and plated covers with EMI gaskets.

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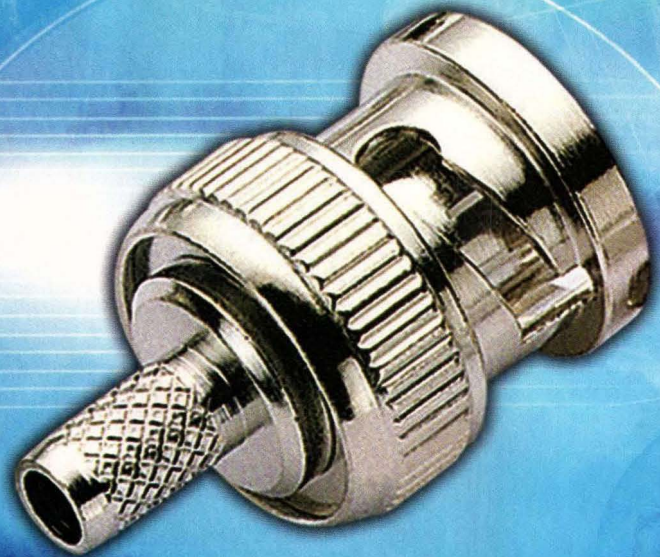
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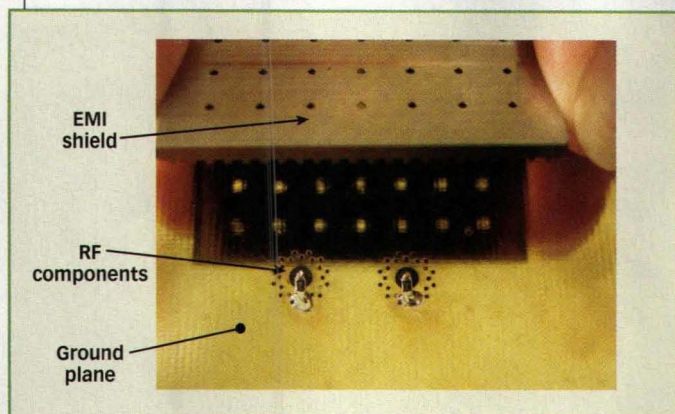
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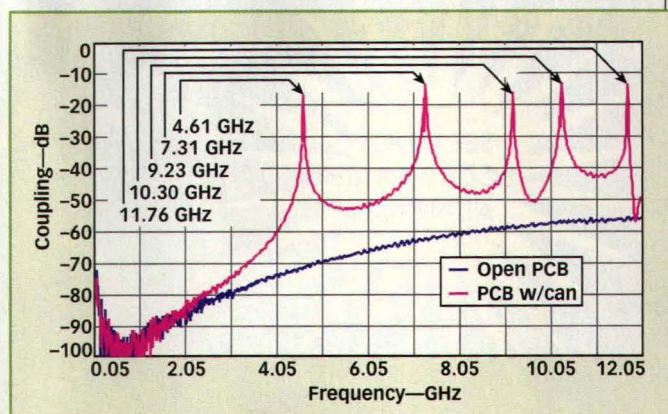
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2. A cavity resonance test fixture was developed to mimic the effect of a metallic cover placed over RF components.



3. The test fixture was used to evaluate coupling with and without a shielding cover, with the formula used for calculations.

mechanical and electrical requirements. Phenomenon such as cavity resonance, aperture radiation, and planar shielding are factors RF engineers face when designing shielding enclosures. The problem is further complicated by the fact that accurate EM field prediction from complicated PCB assemblies, particularly in the near field, is virtually impossible,

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To create the Faraday cage required for proper shielding, a metallic enclosure must be placed around and in close proximity to the components on a PCB. Unfortunately, this may have adverse effects on the performance of the components and the functionality of the cir-

cuit, with the greatest concern being enclosure (cavity) resonances at any of the PCB's operating frequencies. To study this, a simple test fixture was designed to mimic the effect of placing a metallic enclosure over RF components.

The test fixture (Fig. 2) consists of two 50- Ω , 0805 resistors that are launched from SMA connectors from the oppo-

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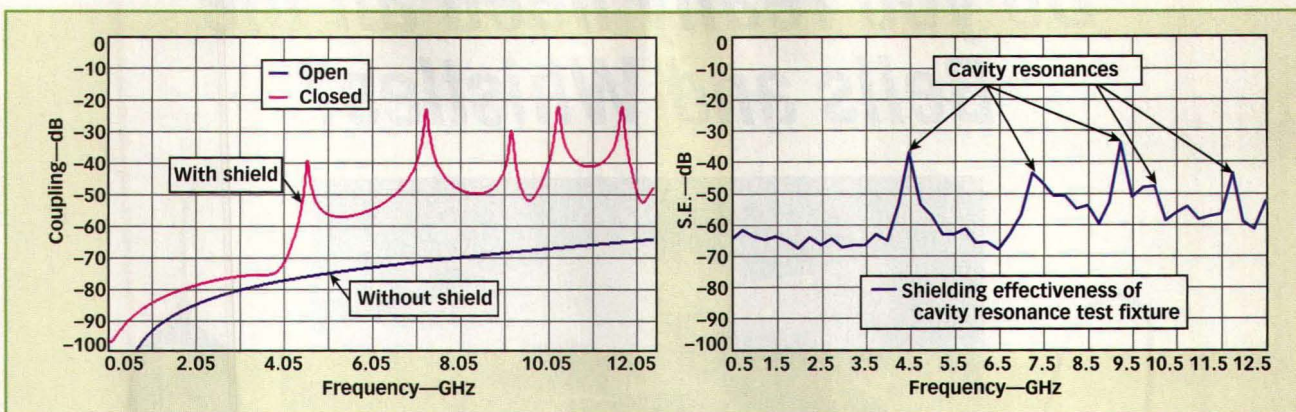
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4. EM field software was used to predict the effects of PCB shielding with and without a cover.

site side of the ground plane. The spacing is arbitrarily set at 0.5 in. (1.27 cm) so that minimal coupling would occur between the two components when the shield is not in place (Fig. 3). Coupling was determined from $20\log S_{21}$ measurements on a microwave vector network analyzer (VNA). A perforated metal square can, with inside dimensions of 1.805×0.114 in. (4.584×0.29

cm), was soldered over the components to illustrate the effects of a metal cavity. The simple formula in Fig. 3 was used to roughly calculate the resonant modes of the EMI enclosure. The formula applies to rectangular cavities and is fairly accurate if the cavity is filled with air. However, most enclosures on PCBs will include the PCB material and components within them, raising the cavi-

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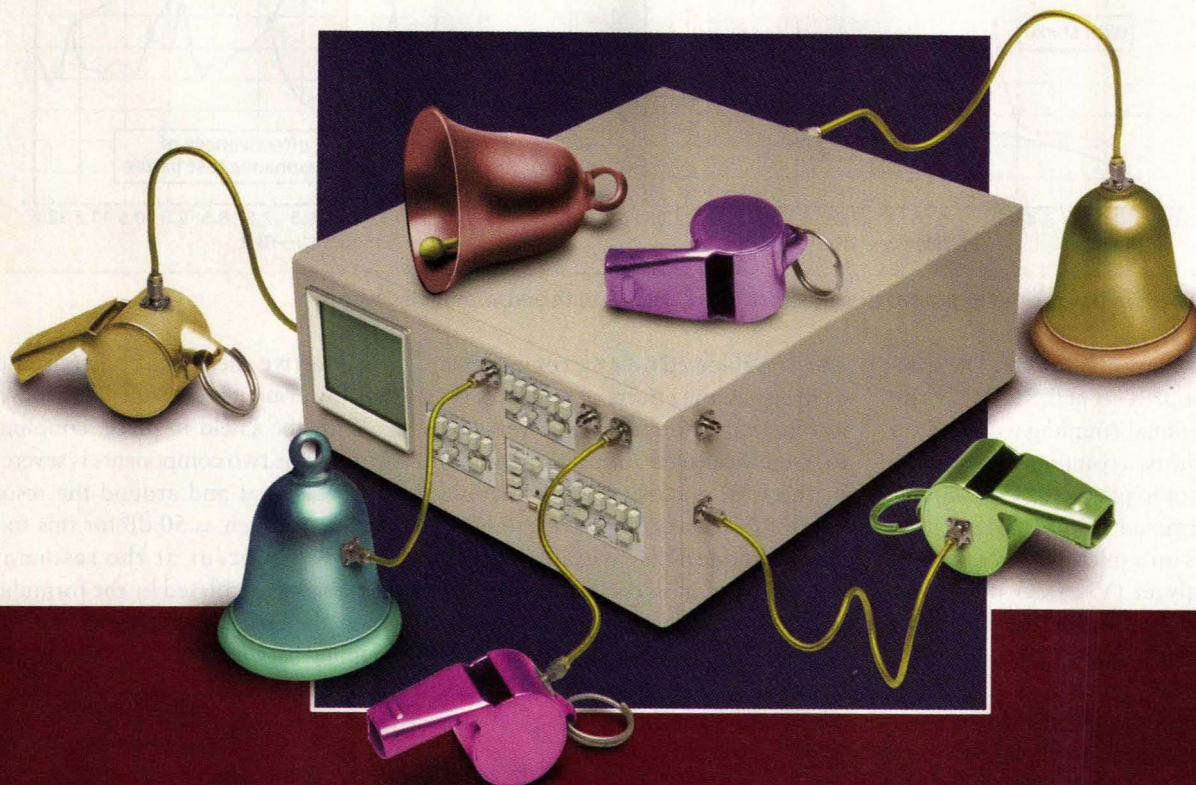


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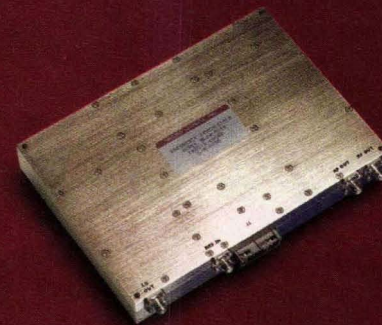
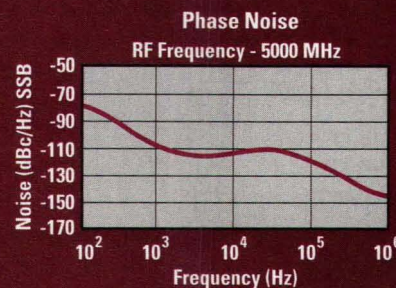
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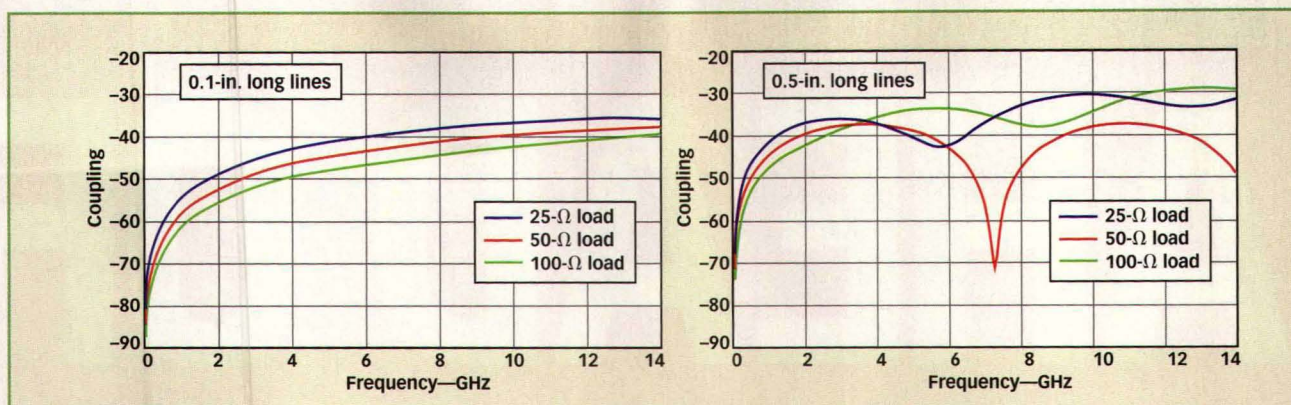


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5. The coupling characteristics of short transmission lines can be used to develop an inefficient radiator for testing SE.

is virtually unchanged. Thus, it is critical to consider these resonant conditions when designing EMI enclosures.

A more accurate way to predict the effects of a rectangular cavity involves the use of an EM field simulator, such as the Sonnet[®] Professional Planar Software Suite from Sonnet Software (Liverpool, NY). This software models planar circuits within a metallic box, and

can easily be used to examine the effects of cavity dimensions, substrate material, and metal wall conductivity. (A free version of the software is available on Sonnet's website at www.sonnetusa.com.)

Figure 4 shows the response of a model produced by the software, with close agreement to the actual measured data.

Another important consideration concerning cavity resonance is the effect

on the shielding performance of the enclosure. Since energy inside the cavity is amplified at the resonant modes, it is likely that the shielding effectiveness (SE) would be lowest at these frequencies. As Fig. 4 shows, the SE drops significantly at each cavity resonance (due to the dimensions of the cavity).

The methods for evaluating the effectiveness of PCB shields can be broken

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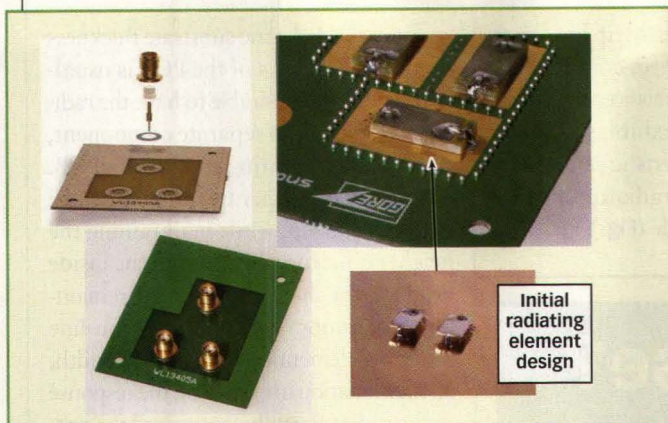
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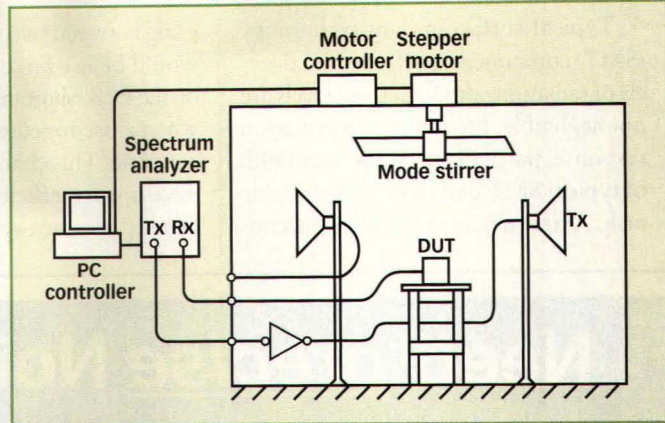


6. This special test fixture includes a custom SMA connector designed for minimal EMI leakage.

into three categories: compliance testing, functional testing, and indirect testing. Compliance testing involves evaluation of the final product given industry-standard test methods and acceptance levels, such as spurious emissions, susceptibility, or electrostatic discharge (ESD). Functional testing involves

evaluation according to performance requirements set by the manufacturer. Intercavity shielding, radiation from the antenna back into the receiver (Rx), and phase noise are examples of parameters that would be considered.

The last category, indirect testing, is used when describing shielding prod-



7. In a mode-stirred shielded chamber, the radiation characteristics of a DUT are compared to those of a reference horn.

ucts. Methods such as MIL G 83528B, ASTM D 4935, and ASTM D 991 are guidelines followed by shielding manufacturers to evaluate their products. These methods can characterize the constituent properties of a shield, but cannot predict the performance in a specific application.

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Typical surface-mount-technology (SMT) components suffer minimal levels of radiation, although these levels are not negligible. Predicting the radiation response, particularly in the near field, of typical SMT devices is a formidable task. A first step is to consider the cou-

pling between two microstrip lines that would be used to connect components on the PCB. Short transmission lines, even when close together, exhibit very little coupling. This characteristic can help to develop an inefficient radiator that can be used as a test vehicle (Fig. 5).

Since the dielectric substrate thickness on the outer layers of the PCB is usually very thin, it is desirable to have the radiating element be a separate component, instead of integrating it onto the PCB. The height of a separate radiator can be readily adjusted, while maintaining the height sufficiently lower than the inside height of the shield. By using the remaining dimensions of the transmission-line radiating element (height, length, width, and termination impedance), the response of the radiator can be optimized for testing shields on PCBs.

The special test fixture developed for testing shields on PCBs (Fig. 6) includes a custom-designed SMA connector that, using a solder pre-form, completely seals the launch point to the ground plane of the PCB. The connectors then feed the radiating elements that are attached to the opposite side of the PCB. The shields would then be centered and attached over each of the elements.

The mode-stirred reverberation chamber technique (Fig. 7) is an excellent EMI test method because of its high dynamic range and repeatability.^{4,5} In this technique, the radiation characteristic from the device under test (DUT) is compared to that of a reference horn antenna. Measurements are first performed on the horn antenna, then the DUT is substituted for the horn and tested. Of course, only one radiator/cavity can be tested at a time. The radiator is first measured without a shield over it, then the shield is attached and the device is re-measured. The SE is calculated as the difference between the received power levels (in decibels) before and after the shield is applied.

The frequency range of such a test is determined by the room dimensions, the test equipment, and the antenna bandwidth. The main limitation is usually the lower frequency boundary, determined by the room size and antenna used. A frequency range of 1 to 13 GHz was used for the tests in this article.

Practically, EMI enclosure cannot be made to be a complete Faraday shield. Gaps due to perforations in the shields, incomplete shields, breaks in the shielding gasket, spaces between ground-

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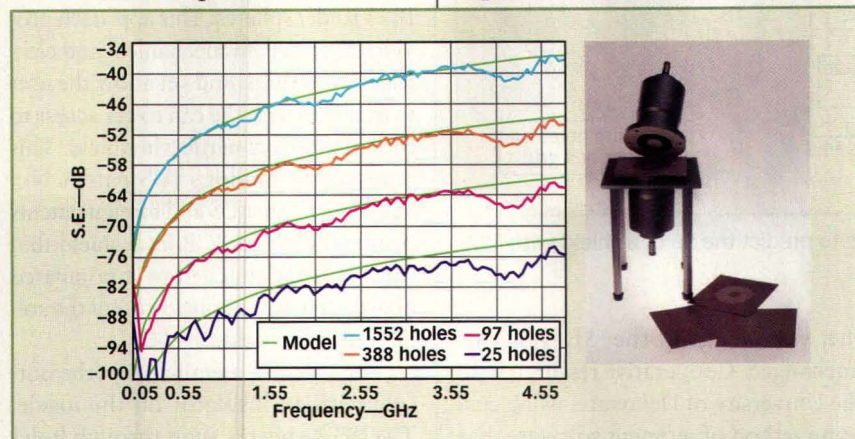
ing vias, and relief areas in the ground plane are necessary to manufacture the complete PCB. But as long as the size of the aperture is much smaller than the wavelength of the highest frequency of interest, it should not cause an appreciable amount of leakage.

The effects of apertures much small-

er than a wavelength at the highest operating frequency of interest has been studied to great lengths.⁹⁻¹¹ For perforated screens, the formula of Fig. 8 has been used to show the frequency relationship between the size of the aperture and SE. Although this frequency dependence represents an accurate rela-

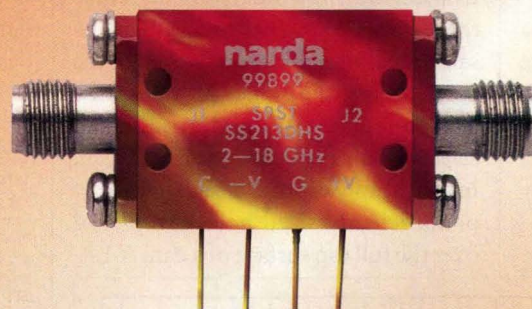
tionship for the far-field response of a large array of holes, an offset factor in SE can deviate quite a bit from real-life applications. To overcome this, a relationship for the far-field (plane-wave) response was derived using empirical data taken from thin copper sheets perforated using the traditional hexagonal pattern. Initially, 1-mm-diameter holes were placed on a 1.7-mm hexagonal grid, such that about 1552 holes fell within the annulus of a typical ASTM D 4935 test cell. The SE was obtained using S_{21} measurements with and without the sheets in place (Fig. 8).

A simple model was generated to represent this test pattern, using the formula of Fig. 8 and a correction offset factor. The next three test patterns were generated by doubling the spacing between the holes each time. This way, about four times fewer holes fell within the annulus of the coaxial cell for each pattern. To generate the mod-



8. Vector network measurements were used to develop a model that relates the size and number of apertures in a perforated screen to the SE of the screen.

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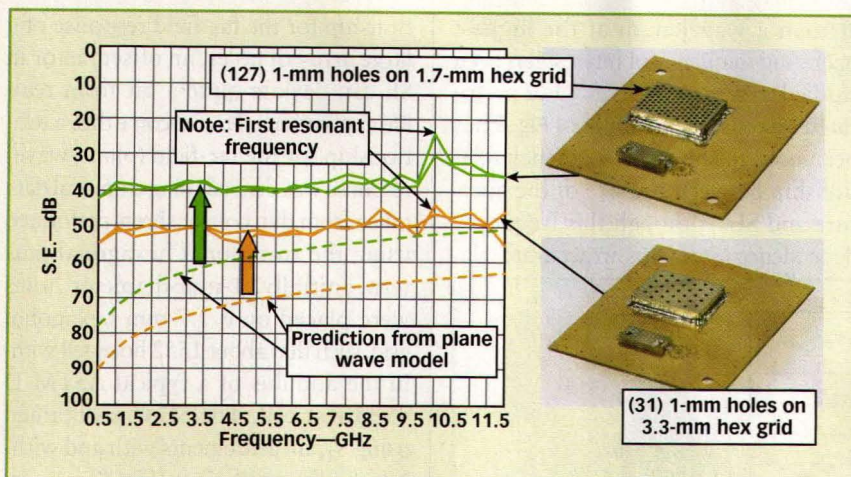
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9. Using the model from Fig. 8, it was possible to predict the SE of a shield with 127 1-mm holes on a 1.7-mm hex grid.

eled data, the original model was changed by a factor of 4 each time, which yielded 12-dB offsets. As Fig. 8 shows, the frequency relationship follows what would be expected for plane-wave excitation of perforated thin sheets.

The next logical step is to see how a perforated soldered can would perform using the same 1-mm hole size on a 1.7-mm grid. Using the test method described earlier, a $16.3 \times 22.5 \times 3.1$ -mm can, completely soldered around its perimeter, was used as the shield. Using the model generated from the plane-wave experiment, a prediction was made for (127) 1-mm holes (Fig. 9). The surprising result of an overall lower SE is added to the fact that the response is flat with frequency, even at low frequencies. This shows that even at low frequencies, where the size of the apertures are extremely small compared to

the wavelength, the SE remains unchanged. Cooperative research with the University of Delaware, using custom method of moment software, has confirmed these results (and will be published formally at the 2003 IEEE EMC symposium in Boston, MA). If the aperture spacing is increased by a factor of 2, thus reducing the total number of holes by a factor of 4, the SE is increased, or shifted by about 12 dB, the same as the plane-wave case.

Perforations in cans allow for heat transfer during solder reflow. Since these cans are attached to a PCB by the same SMT assembly process used for the components, the solder reflow of the components under and around the shields should not be impeded by the can. In typical applications, hexagonal arrays of sufficiently sized holes are placed over the full top surface of a can. A 1.4-

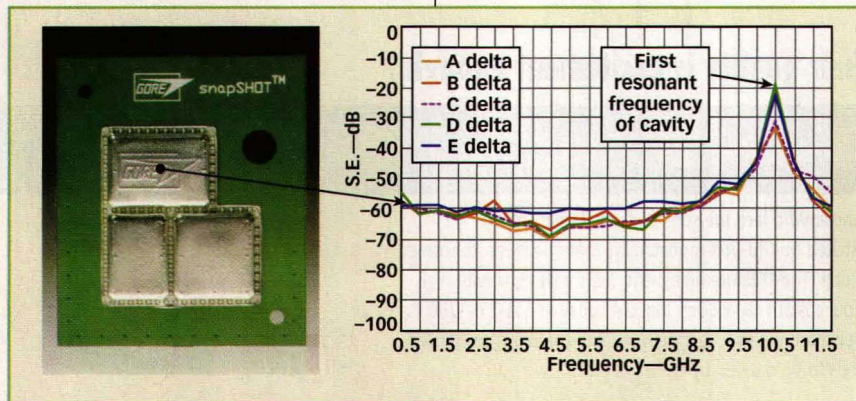
mm or larger hole is typical with the hexagonal grid yielding a hole spacing of 4.4 mm and smaller.

To simplify the addition of shielding to PCBs, Gore has developed the "snap-on" snapSHOT shield,TM which consists of a metallized thermoformed shell that is attached to a PCB using standard BGA solder spheres. This approach provides an excellent mechanical and electrical connection and yet allow the user to easily remove the can to get access to the components within the shield. This patented technology (US patent No. 6,377,475; other US and foreign patents pending) offers a PCB-level shield that can be easily attached to a populated board after it has gone through the solder reflow process.

The shield is metallized on the outside with an insulator on the inside. The BGA spheres snap through holes in the shield, creating a robust electrical and mechanical connection to the PCB. The periodicity of the sphere placement is determined by the required shielding performance. Since the inside surface is not conductive, any components that may come in contact with this surface will not be electrically shorted. The snapSHOTTM shield provides significantly improved shielding over similar perforated solder cans with a great deal of design flexibility (Fig. 10). **MRF**

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10. The snapSHOTTM shielding technology is based on flexible thermoformed materials with high SE at microwave frequencies.

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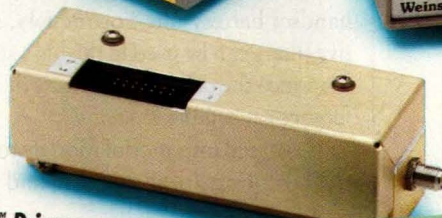
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150T-75	0-75/5		4
150T-110	0-110/10		4
151T-11	0-11/1	dc-4 GHz	4
151T-15	0-15/15		4
151T-31	0-31/1		5
151T-62	0-62/2		5
151T-70	0-70/10		3
151T-75	0-75/5		4
151-110	0-110/10		4
152T-11	0-11/1	dc-26.5	4
152T-15	0-15/1		4
152T-55	0-55/5		4
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Band-reject filters can be applied quite effectively in reducing or removing unwanted IMD interference signals from the operating band of a cellular receiver system.

Interference in cellular receive systems can stem from a variety of causes. Generally, the interference starts at a base-station transmitter (Tx), either from the same cellular system or from a nearby Tx. The interference can result in dropped calls, decreased receiver (Rx) sensitivity (and range), increased Rx noise figure and desensitization of receive-system active components. This last situation can

er-order signal products caused by the nonlinear mixing of two or more transmitted carriers, and is a well-understood

phenomenon. These IMD products can block the reception of desired signals when they fall into the receive band and exceed the level of the desired receive signals (typically -70 to -100 dBm) at the front end of the base-station Rx (**Fig. 1**).

Passive components in the transmit signal path can, and do, create IMD products whenever two or more transmit carriers are present. The key to the severity of the interference is the level of these products relative to the base-station Rx sensitivity; ideally, these IMD products should be limited to a level that is below

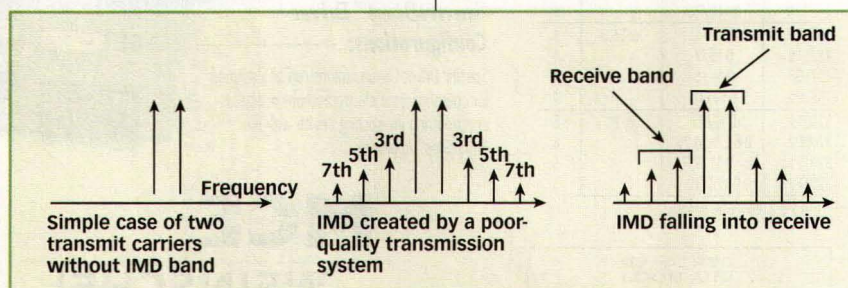
lead to dropping all calls in an individual cellular sector. For code-division-multiple-access (CDMA) handset operators far from a base station (in the far field) with elevated noise floors, the base station will command the mobile units to increase output power. This will lead to higher handset emissions and reduced handset battery life. Fortunately, selective filters can be used in the base station to reduce the deleterious effects of interference on cellular systems.

Tx-based intermodulation distortion (IMD) consists of third, fifth, and high-

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1. Two ideal transmit carrier signals (left) generate IMD due to the nonlinear effects of transmit components. The IMD can fall into a receiver's operating band and degrade receive performance.

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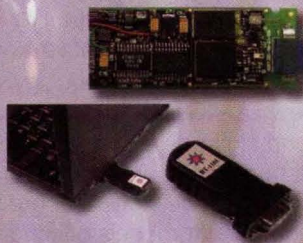
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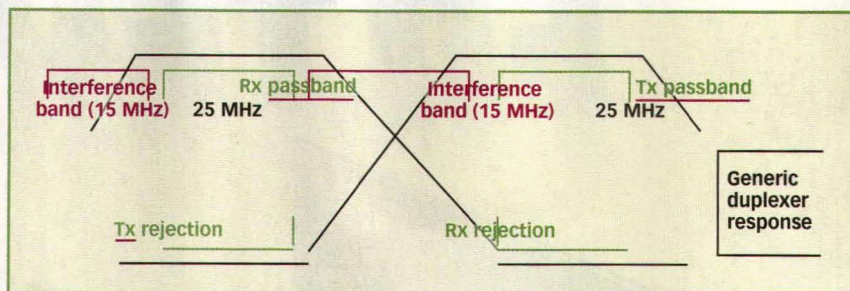
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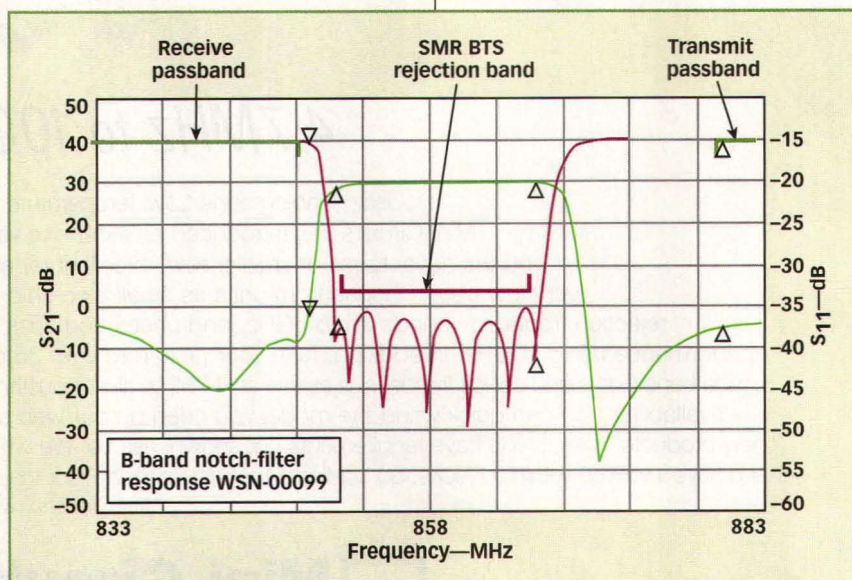
2. This generic duplexer has been optimized for transmit-band rejection, allowing interfering signals to pass unrejected into the receiver-operating band.

the usable sensitivity threshold of the receive system. Therefore, the integrity of a cellular Tx system is the level of IMD products when multiple carriers are present.

This type of interference to the Rx system can only be solved by improving the intermodulation characteristics of the Tx signal-path components, or by filtering the output of the Tx with a low IMD filter. Since the filter increases signal loss and affects the downlink signal budget, the preferred approach is to control IMD in the transmit-path components, such as antennas, filters/duplexers, power amplifiers (PAs), combiners, cables, connectors, couplers and any other passive or active components in the signal path. IMD in the passive components can be limited through the use of high-quality silver (Ag) or gold (Au)-plated contacts; nickel (Ni)-plated or passivated contacts

should be avoided. Tx components should be specified in terms of maximum allowable IMD under specified conditions. Once IMD radiates though the antenna into free space, it cannot be filtered from the receive system without also attenuating the levels of desired signals.

Rx desensitization occurs when a co-located or nearby Tx's high signal strength enters an Rx's signal chain and causes IMD that interferes with or completely blocks the reception of the intended low-level signals. At high enough levels, this IMD can cause the Rx's active components to enter saturation, increasing the noise floor of the Rx's front end low-noise amplifier (LNA) dramatically. In CDMA cellular systems, this type of interference will also cause a base station to send commands that increase the transmit power of far-field handsets (to overcome the rise in noise floor caused by the IMD),

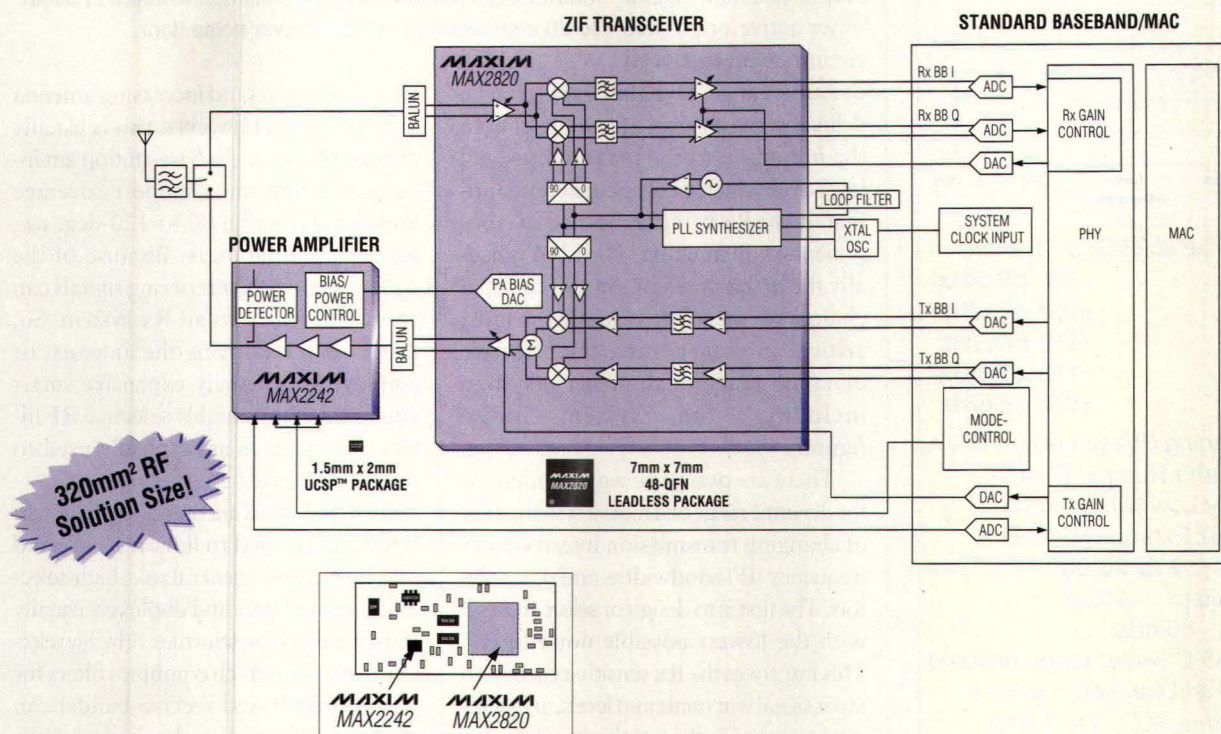


3. Notch filters for cellular-band frequencies allow the rejection of interference from co-located SMR transmitters.

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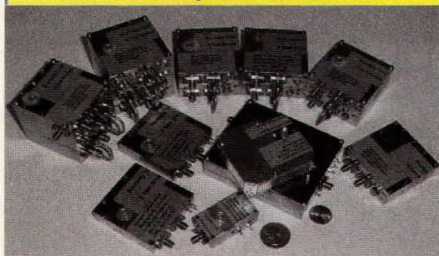
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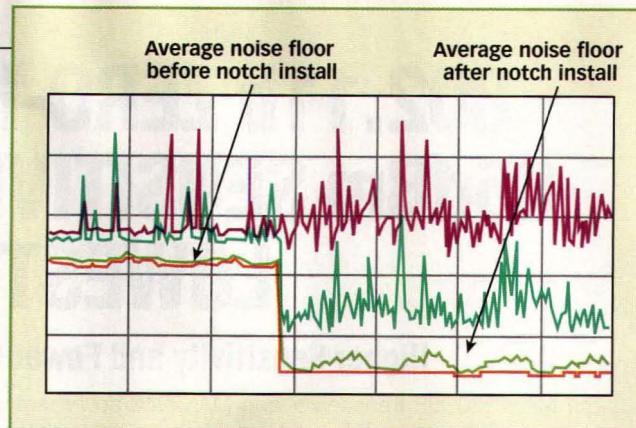
DESIGN

leading to higher handset emissions and greatly reducing handset battery life.

Fortunately, there are several solutions for this second type of IMD interference. The first solution is to ensure that the active Rx stages use amplifiers, mixers, and other active or passive circuitry with enhanced dynamic range. Dynamic range can be defined as the difference (in decibels) from the minimum to the maximum signal level over which a component (and ultimately the Rx being composed of components) will function. The LNA is usually the first active stage in an Rx signal chain and, therefore, usually the most critical in determining the effective dynamic range of the receive system including the system noise figure.

There are two basic ways to increase the dynamic range of a receive system, short of changing transmission intermediate-frequency (IF) bandwidths and other factors. The first is to design or select an LNA with the lowest possible noise figure. This improves the Rx sensitivity when an input signal is at minimum levels, and effectively improves the sensitivity of the Rx system for far-field handset users. The second method is to increase the third-order intercept point (IP3) of the LNA. The IP3 specification is a figure of merit for an amplifier's ability to handle high signal levels without odd-order IMD products (third, fifth, etc.) increasing beyond an acceptable level, and falling into the receive band, causing receive system interference. By increasing an LNA's IP3 characteristics, the receive system's dynamic range increases for higher-level signals. In general, designers select LNAs with as much dynamic range (lowest noise figure and highest IP3) as possible within cost and power budgets.

The second path to improving Rx desensitization is by limiting the out-of-band or interfering signals from reaching the Rx. This can be accomplished by decreasing the beamwidth of the base-



4. The addition of a dual-notch filter assembly in a major metropolitan B-band cellular installation resulted in about 30-dB improvement in the receiver noise floor.

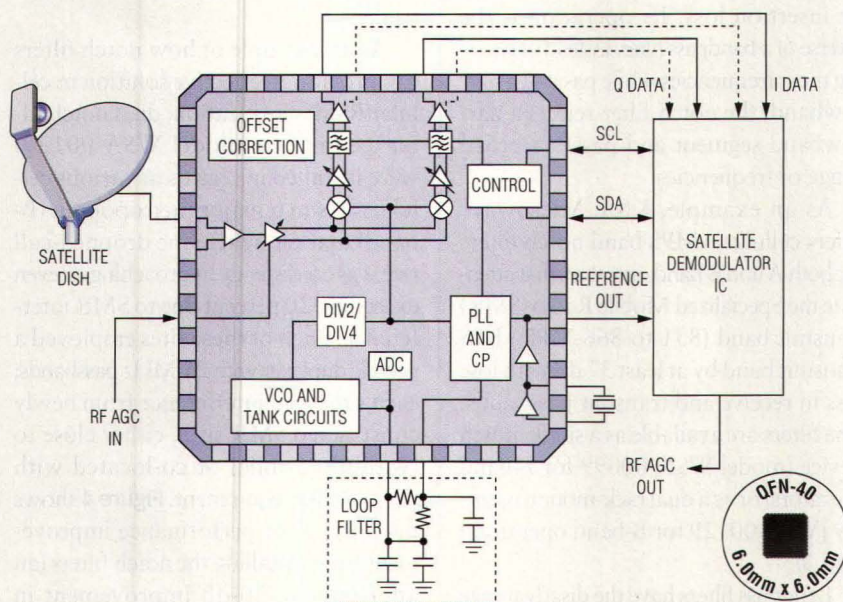
station antenna (and increasing antenna and link gain). However, this is usually not possible since the base-station antenna beamwidth must support coverage within a (typically) 60-to-120-deg. sector for multiple users. Because of the wide beamwidth, interfering signals can easily gain entry to an Rx system. So, rather than modifying the antenna, or employing extremely expensive smart antenna systems, highly selective RF filters or duplexers are typically used to reject IMD-based interference.

Both bandpass and band-reject (notch) filters can be used to limit receive-band interference. As a general rule, high-selectivity receive filters and duplexers usually employ a bandpass structure. A highly selective duplexer (which combines filters for both transmit and receive bands) can provide high receive-path rejection above and below the passband of interest, thus rejecting by a specified level potential or existing interference signals. Because of antenna site/zoning constraints, duplexers are attractive since they allow antennas to be consolidated or reduced at a given site. Because of zoning restrictions, most sites require duplexers for receive and transmit signals. High-performance duplexers offer a suitable solution for new sites or those undergoing major upgrades. For sites already employing "generic" duplexers, which have operated successfully for many years, newly installed or co-located competitive TxS can unveil the shortcomings of these generic duplexers. For example, if a duplexer has been optimized for transmit rejection, and uses a wide receive filter bandwidth with inadequate rejection of IMD interference, the Rx can become

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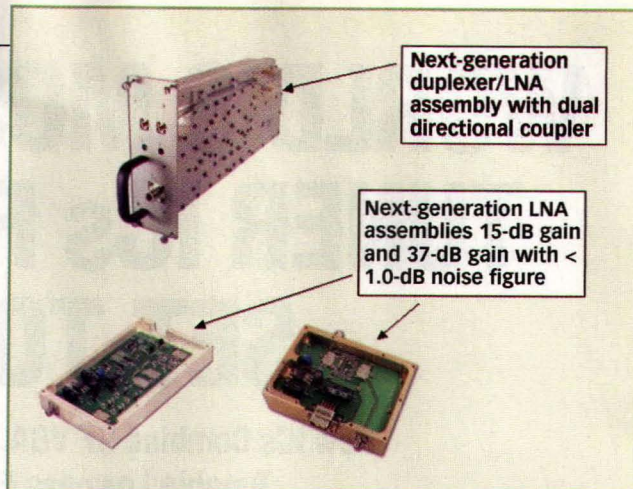
DESIGN

desensitized by sufficiently high-level IMD interference signals that are allowed to pass through the duplexer (Fig. 2).

Fortunately, a band-reject or notch filter offers a novel solution to Rx desensitization. Although available for decades, the application of this type of filter in cellular systems has been limited to extreme spot interference or test-system requirements. The unique advantage of the notch filter is its ability to reject only the specific band of interest, and pass the intended receive and transmit bands with very little insertion loss. Its operation is the inverse of a bandpass filter. Instead of rejecting most frequencies while passing a narrowband, the notch filter rejects a narrowband segment and passes a broad range of frequencies.

As an example, K&L Microwave offers cellular AMPS band notch filters for both A and B Band operators that attenuate the Specialized Mobile Radio (SMR) transmit band (851-to-866-MHz) BTS transmit band by at least 37 dB with low loss in receive and transmit passbands. The filters are available as a single-notch device (model WSN-00099 for B-band operators) or as a dual rack-mount assembly [WSA-00129 for B-band operators] (Fig. 3).

Bandpass filters have the disadvantage of affecting the isolation of the receive system relative to one's own Tx. As a result, extensive re-engineering and re-specification may be required for a system in which bandpass filters are used to meet the original equipment manufacturers' (OEMs) transmit rejection. The notch-filter solution rejects only the interfering band of interest, and allows the otherwise properly functioning generic duplexer and cellular transceivers to remain in place. The notch-filter approach offers very low signal loss in both receive and transmit passbands, and allows for "in-line" installation in both duplexed or receive-only antenna feeds. For duplexed sys-



5. This integrated assembly includes two LNAs, a dual directional coupler, and high selectivity cellularband duplexer for limiting out-of-band interference signals.

tems where the transmit-signal levels can be high, the notch filters mentioned above have been designed and tested to handle very high RF power levels, to 500 W CW and 10 kW for instantaneous peaks.

As an example of how notch filters can provide an effective solution to cellular Rx desensitization, dual notch filter assemblies (model WSA-00129) were installed in sectors at various cellular sites in a major metropolitan (B-band) market, where the dropped-call rate was consistently approaching or even exceeding 20 percent due to SMR interference. Each of these sites employed a generic duplexer with 25-MHz passbands; each site faced interference from newly constructed SMR sites, either close to (within 0.5 mile) or co-located with the providers equipment. Figure 4 shows the noise-floor performance improvement from installing the notch filters (an approximate 30-dB improvement in the noise floor). Dropped calls improved to less than 2 percent.

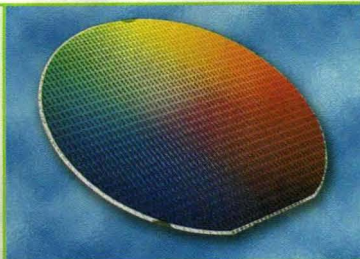
Another approach to the cellular Rx desensitization problem is the use of high-performance LNAs integrated with a selective receive-band filter. An example of this type of assembly (from K&L Microwave) employs high IP3 LNA and an integrated duplexer to achieve 2-dB noise figure and >30-dB gain with less than 5-MHz passband and tunability over the full AMPS cellular band (Fig. 5). One of the LNAs features 15-dB gain and output IP3 of +37 dBm while the higher-gain LNA offers 37-dB gain and +42 dBm output IP3. MRF

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Analyze Coax Cable Digital Pulse Distortion

This analysis shows the effects of a short length of coaxial cable and associated test equipment on the characteristics of a high-speed digital pulse.

Coaxial cables are often neglected during the characterization of high-speed digital components, even though such transmission lines can have a major electrical influence on a test setup. Coaxial cables are sometimes treated as lossy elements, and assigned a lumped capacitance and/or inductance, although the electrical effects of a coaxial cable can be much more complex than a single capacitance value.

pulse. The Fourier Transform, and inverse Fourier Transform for the pulse are¹

$$V(f_k) = \frac{1}{N} \sum_{i=0}^{N-1} V(t_i) e^{-j \frac{2\pi k}{N} i}$$

$$V(t_n) = \sum_{i=0}^{K-1} V(f_i) e^{j \frac{2\pi n}{K} i} \quad (1)$$

where:

$$M = 17,$$

$$N = 2^M,$$

$$K = 1 + 2^{M-1}$$

$$t_n = \Delta t (-2^{M-1}),$$

$$\Delta t = 0.25 \text{ ps},$$

$$n = 0 \dots (N-1)$$

$$f_k = k f_0,$$

$$f_0 = 1/(2^M \Delta t), \text{ and}$$

$$k = 0 \dots (K-1). \text{ Note that the values}$$

for M and Δt have been chosen as required for optimal resolution and bandwidth.

At the input, this pulse is very narrow and has a finite rise time. It can be expressed mathematically as:

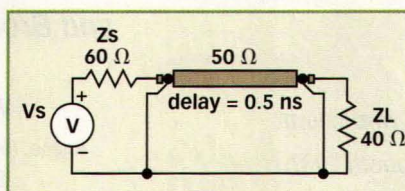
SEE EQ. 2 ON P. 98

where:

A = amplitude,

What follows is an examination of how the electrical performance of a coaxial cable, notably loss and distributed capacitance/inductance, can affect the integrity of a high-speed digital pulse. The article will review the creation of a high-speed digital pulse, how to launch the pulse from an ideal output impedance, follow its transmission down an imperfect coaxial cable, and show how to make measurements at a perfect load.

To apply transmission-line theory to this analysis, the pulse will be created in the time domain and then converted to the frequency domain. To use the analysis tools available, the pulse is created as a discrete time



1. This setup can be used to verify the effectiveness of transmission-line theory.

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S4W2	S4W5	N4W5	4	±0.40
S5W2	S5W5	N5W5	5	±0.40
S6W2	S6W5	N6W5	6	±0.40
S7W2	S7W5	N7W5	7	±0.60
S8W2	S8W5	N8W5	8	±0.60
S9W2	S9W5	N9W5	9	±0.60
S10W2	S10W5	N10W5	10	±0.60
S12W2	S12W5	N12W5	12	±0.60
S15W2	S15W5	N15W5	15	±0.60
S20W2	S20W5	N20W5	20	±0.60
S30W2	S30W5	N30W5	30	±0.85
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τ = the pulse width, and
 t_r = the 100-percent rise time,
 and

SEE EQ. 3 AT RIGHT

Transmission-line theory is well established and this analysis should be straightforward. Still, it may provide some reassurance by running a simple analysis that can be confirmed by means of time-domain-analysis techniques. In this case, a pulse will be launched from a non-ideal source and transmitted a short distance to a nonideal load. The source and the load are connected by means of a short, lossless coaxial cable. The source is chosen to be 60 Ω , the cable is 50 Ω , and the load 40 Ω (Fig. 1). These impedance mismatches will generate multiple reflections, which can be calculated using time-domain-to-frequency-domain-to-time-domain theory, and then verified using time-domain

$$V_s(t_n) = \frac{A\tau}{2t_r} \left[1 - \frac{|t_n - \tau/2|}{\tau/2} \right] \Pi \left(\frac{t_n - \tau/2}{\tau} \right) - \left(\frac{A\tau}{2t_r} - A \right) \left[1 - \frac{|t_n - \tau/2|}{\tau/2 - t_r} \right] \Pi \left(\frac{t_n - \tau/2}{\tau - 2t_r} \right) \quad (2)$$

$$\Pi \left(\frac{t}{2\tau} \right) = u(t + \tau) - u(t - \tau) \quad (3)$$

$$H_{ABCD}(f) = \begin{pmatrix} H_A & H_B \\ H_C & H_D \end{pmatrix} = \begin{pmatrix} \cos \beta l & jZ_0 \sin \beta l \\ \frac{j}{Z_0} \sin \beta l & \cos \beta l \end{pmatrix}$$

βl = cable electrical length

$$= 2\pi f(0.5 \text{ ns}) \quad (4)$$

$$S_{11}(f) = \frac{Z_L H_A + H_B - Z_S Z_L H_C - Z_S H_D}{Z_L H_A + H_B + Z_S Z_L H_C + Z_S H_D} \quad (5)$$

transmission-line theory.

The analysis will be presented as a time-domain graph of the source voltage, the input voltage, and the output voltage. The first step in the analysis is to convert the source voltage into the frequency domain by means of the Fourier Transform. In order to calculate the input volt-

age, the input impedance, which can be determined by the scattering parameter (S parameter) S_{11} , is required. The S_{11} term can be calculated from the circuit's ABCD parameters²:

SEE EQ. 4 ABOVE

where:

βl = the electrical length of the cable
 $[= 2\pi f(0.5 \text{ ns})]$.

From the ABCD parameters, S_{11} is³

SEE EQ. 5 ABOVE

The input impedance is:

$$Z_{in}(f) = Z_S \left(\frac{1 + S_{11}}{1 - S_{11}} \right) \quad (6)$$

The input voltage and current become:

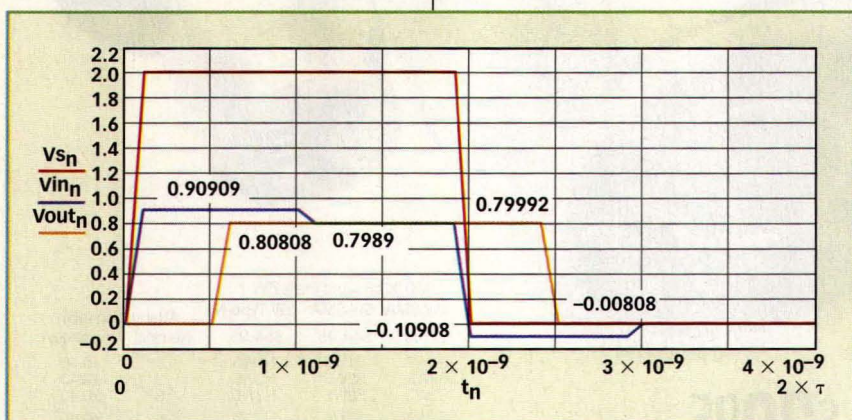
$$V_{in}(f) = V_s(f) \frac{Z_{in}(f)}{Z_{in}(f) + Z_S}$$

$$I_{in}(f) = \frac{V_{in}(f)}{Z_{in}(f)} \quad (7)$$

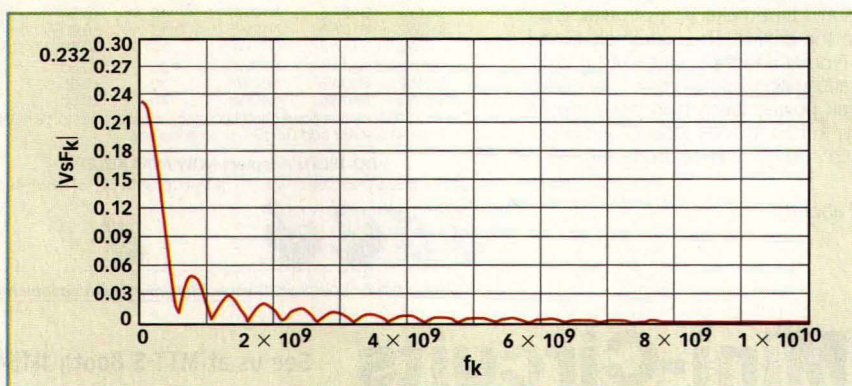
The output voltage and current can be calculated using the ABCD parameters and the input voltage and current:

$$\begin{bmatrix} V_o(f) \\ I_o(f) \end{bmatrix} = (H_{ABCD})^{-1} \begin{bmatrix} V_{in}(f) \\ I_{in}(f) \end{bmatrix} \quad (8)$$

Finally, the input and output voltages are transformed back into the time domain. Graphs of the source, input, and output voltages from this analysis are



2. These source, input, and output voltages are checked by transforming back and forth between the time and frequency domains.



3. This is the frequency response of the source voltage feeding the coaxial cable.

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AFS4-02001800-50-LN	2 - 18	18	2.50	5.0	2:1	+10	\$ 995	\$ 895
ULTRA-WIDEBAND								
AFS3-00100200-18-LN	.1 - 2	36	1.00	1.8	2:1	+10	\$ 750	\$ 675
AFS3-00100400-22-LN	.1 - 4	28	1.25	2.2	2:1	+10	\$ 750	\$ 675
AFS3-00100600-25-LN	.1 - 6	24	1.50	2.5	2:1	+10	\$ 750	\$ 675
AFS3-00100800-32-LN	.1 - 8	24	1.50	3.2	2:1	+10	\$ 750	\$ 675
AFS3-00101000-38-LN	.1 - 10	20	1.50	3.8*	2:1	+10	\$ 750	\$ 675
AFS3-00101200-42-LN	.1 - 12	20	1.75	4.2*	2:1	+10	\$ 750	\$ 675
AFS4-00101800-55-LN	.1 - 18	20	2.50	5.5*	2.5:1	+10	\$ 900	\$ 810
AFS4-00102000-60-LN	.1 - 20	16	2.75	6.0*	2.5:1	+10	\$ 995	\$ 895

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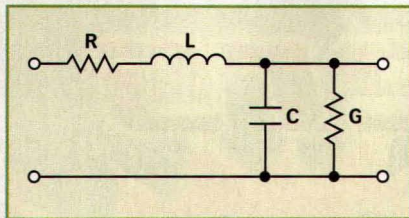
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4. This distributed-element circuit model represents the coaxial cable and its connectors.

shown in Fig. 2. The rise times (10 to 90 percent) of each signal are as follows: $t_r^s = 80$ ps, $t_r^{in} = 80$ ps, and $t_r^L = 80$ ps at the output. The fact that all three rise times are the same is to be expected, since the source and load impedances are real and the cable connecting the source and load is without loss. But, although the cable is considered without loss, it nonetheless has distributed inductance and capacitance (see Fig. 3 and Eqs. 23 and 24). In the frequency response of the source voltage, it should be noted that the finite rise time changes the frequency spectra from that of an ideal pulse. Because the coaxial cable is lossless, the time-domain responses of the input and load voltages can be calculated using the reflection coefficients and the delay time of the coaxial line.

For a coaxial line with rise time of zero, the analysis is as follows. The reflection coefficients at the source and load, respectively, are:

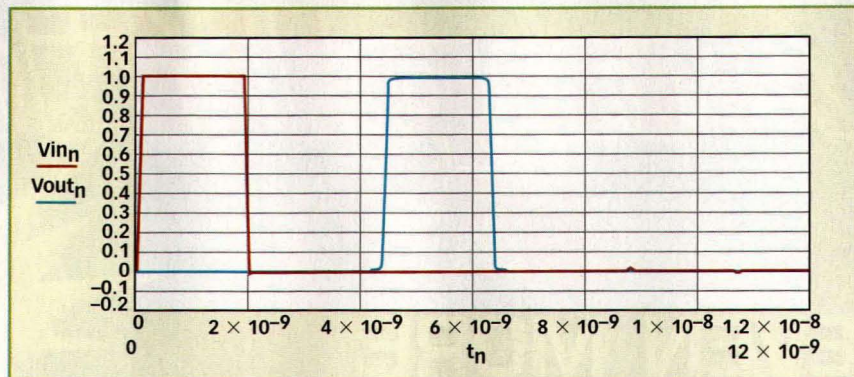
$$\Gamma_s = \frac{Z_s - Z_o}{Z_s + Z_o} = \frac{60 - 50}{60 + 50} = \frac{1}{11} \quad (9)$$

SEE EQ. 10 ON P. 102

At $t = 0$ seconds, a traveling wave of 2-V amplitude is incident on the input of the circuit. The voltage is immediately present at the input, and propagates toward the load. The input voltage, and the wave traveling toward the load (V^+_1), are:

SEE EQ. 11 ON P. 102

The wave V^+_1 travels the length of the transmission line in 0.5 ns. At $t = 0.5$ ns, the traveling wave arrives at the load. Because of the impedance mismatch, a reflected voltage wave (V^-_2) is generated which propagates toward



5. This shows a digital pulse propagated through a nonideal coaxial cable.

the input. The voltages at the load and the reflected wave are:

SEE EQ. 12 ON P. 102

SEE EQ. 13 ON P. 102

Wave V^-_2 travels to the input. Upon arriving at the input ($t = 1$ ns), another reflected wave is generated because of the impedance mismatch of the source.

SEE EQ. 14 ON P. 102

SEE EQ. 15 ON P. 102

At $t = 1.5$ ns, wave V^+_3 arrives at the load, and generates a reflected wave V^-_4 . The reflected wave and load voltage are:

SEE EQ. 16 ON P. 102

SEE EQ. 17 ON P. 104

At $t = 2$ ns, wave V^-_4 arrives at the input. At the same time, a -2 -V traveling wave from the source is incident on the input. The resulting wave, which travels toward the load, and the voltage at the input are:

SEE EQ. 18 ON P. 104

SEE EQ. 19 ON P. 104

SEE EQ. 20 ON P. 104

At $t = 2.5$ ns, wave V^+_5 arrives at the load. The reflected wave and voltage at the load are:

SEE EQ. 21 ON P. 104

SEE EQ. 22 ON P. 104

Reflections continue endlessly with

decreasing amplitude, although no further reflections will be calculated for the sake of this analysis. Comparison with the time-to-frequency-to-time-domain technique shows the two techniques to be equal.

As this analysis shows, a coaxial cable can be approximated as a length of cable with characteristic impedance of 50Ω terminated with two connectors of higher or lower characteristic impedance.⁴ Using this as a foundation, loss terms are added giving a reasonable model for a lossy cable with a nonzero reflection coefficient (VSWR). For the purpose of this analysis, the coaxial cable selected for study is designed for mode-free operation to 35 GHz.

For formulating the S-parameters, the distributed circuit elements for each section of the cable will be calculated, combined into ABCD parameters, and finally converted to S-parameters to evaluate the VSWR and loss. The distributed element circuit model for this purpose is shown in Fig. 4.

The cable is broken into three sections, with characteristic impedances of 50Ω for the cables and 54Ω for the connectors. The cable, which is 36 in. long, has a dielectric constant of 2. The ratio of b/a in the cable is 3.25. The remaining parameters were chosen to produce the characteristic impedance and acceptable VSWR and loss curves. Omitted dimensions are a_i , b_i , and len_i , and electrical parameters ϵ_r , $\tan \delta$, and σ_c^2 :

SEE EQ. 23 ON P. 104

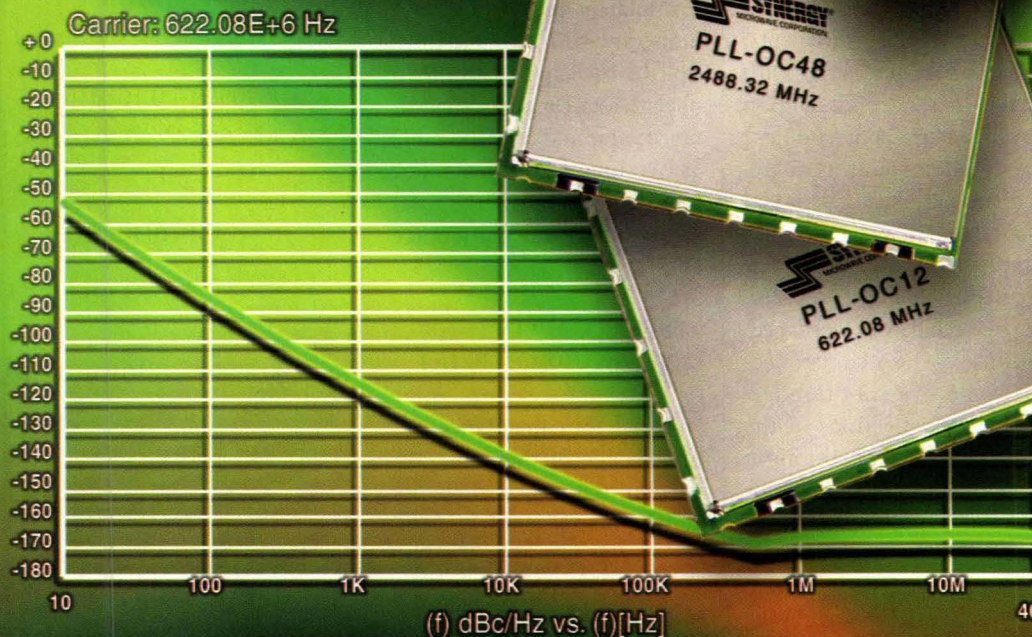
The exact characteristic impedance, and the propagation constant are:

SEE EQ. 24 ON P. 104

SAW

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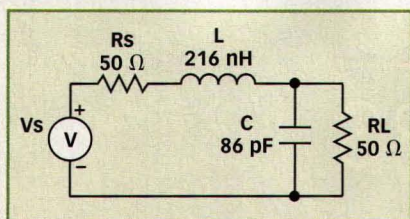
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6. This model represents the capacitance of a coaxial cable along with an ideal load.

The ABCD matrix for the cable is the combination of the ABCD parameters for the individual sections³:

SEE EQ. 25 ON P. 104

The two S-parameters of interest are S_{11} and S_{21} , calculated respectively in Eqs. 26 and 27:

SEE EQ. 26 ON P. 104

SEE EQ. 27 ON P. 104

As a quick check, the frequency spacing between peaks can be calculated as⁴:

SEE EQ. 28 ON P. 104

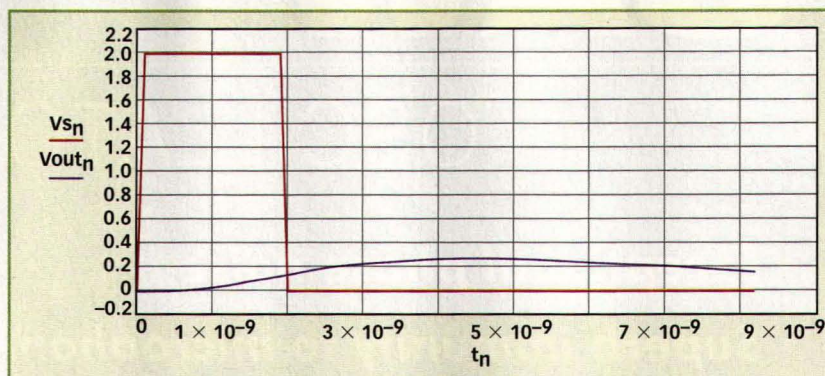
which shows 17 peaks between grid lines, in agreement with measurements. The measured VSWR may differ with the calculated values, depending upon the number of measurement samples.

The primary interest in this analysis is the distortion effects of a coaxial cable on a digital pulse. When the pulse of Fig. 2 is applied to the experimental cable, the output of Fig. 5 results. Although the source and load impedances are 50 Ω, because of the 54-Ω connectors there is some overshoot/undershoot and multiple reflections of the voltage V_{in} . In the case where the input is a pulse train, the multiple reflections would interfere with subsequent pulses. The exact interference would depend on the length of the cable.

While there is significant loss in the cable, the pulse is not reduced because of the pulse width. The 2-ns pulse is primarily contained in the first 500 MHz of bandwidth where the loss is lower. The rise times of the input and output pulses are given by:

SEE EQ. 29 ON P. 104

while the propagation delay of the cable is given by:



7. This plot shows the source voltage and voltage across a perfect load for a lumped-element capacitance/inductance cable representation.

SEE EQ. 30 ON P. 104

Note that there was no change in rise time observed in the lossless cable (Fig. 2). When loss is added to the cable model, however, the rise time increases, demonstrating that increased rise time at the output of a cable is the result of cable losses.

It is tempting to take the formula for capacitance and inductance (Eq. 23) and multiply by the length of cable to obtain a total capacitance/inductance value and treat the cable as two lumped elements. For the cable described, the total capacitance and inductance are:

SEE EQ. 31 ON P. 104

SEE EQ. 32 ON P. 104

When the capacitance and inductance of the cable are treated as lumped elements and the loss and length ignored, the circuit becomes the one shown in Fig. 6. Applying the pulse of Fig. 2, the voltage from the source and that appearing across the load are described by Fig. 7. Clearly, the cable cannot be treated as a lumped capacitance and inductance. Comparing Figs. 5 and 7, it is apparent that treating the cable as a lumped-element capacitance/inductance significantly alters the signal from that which would actually occur.

$$\Gamma_L = \frac{Z_L - Z_o}{Z_L + Z_o} = \frac{40 - 50}{40 + 50} = \frac{-1}{9} \quad (10)$$

$$V_{in} = V_s \frac{Z_o}{Z_o + Z_s} = 0.9090909V \quad 0 < t < 1 \text{ ns}$$

$$V_1^+ = 0.9090909V \quad (11)$$

$$\begin{aligned} V_2^- &= V_1^+ \Gamma_L \\ &= 0.9090909 \times \frac{-1}{9} \\ &= -0.1010101V \quad (12) \end{aligned}$$

$$\begin{aligned} V_L &= V_1^+ + V_2^- \\ &= 0.9090909 - 0.1010101 \\ &= 0.8080808V \quad 0.5 \text{ ns} < t < 1.5 \text{ ns} \quad (13) \end{aligned}$$

$$\begin{aligned} V_3^+ &= V_2^- \Gamma_s \\ &= -0.1010101 \times \frac{1}{11} \\ &= -0.009182736 \quad (14) \end{aligned}$$

$$\begin{aligned} V_{in} &= V_1^+ + V_2^- + V_3^+ \\ &= 0.9090909 - 0.1010101 - 0.009182736 \\ &= 0.7989V \quad 1 \text{ ns} < t < 2 \text{ ns} \quad (15) \end{aligned}$$

$$\begin{aligned} V_4^- &= V_3^+ \Gamma_L \\ &= -0.009182736 \times \frac{-1}{9} \\ &= 0.001020304V \quad (16) \end{aligned}$$

It is not a simple matter to calculate the losses for a pulse and the increase in rise time due to a coaxial cable. The pulse consists of many separate frequency components, each with a different loss value. As these components increase in frequency, so does the loss. Because the edges of a pulse are formed by higher-frequency components, there is an increase in rise time and a rounding of



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$$\begin{aligned} V_L &= V_1^+ + V_3^+ + V_4^- \\ &= 0.9090909 - 0.1010101 - 0.009182736 + 0.001020304 \\ &= 0.79992 \text{ V} \quad 1.5 \text{ ns} < t < 2.5 \text{ ns} \quad (17) \end{aligned}$$

$$\begin{aligned} V_{5A}^+ &= V_4^- \Gamma_S \\ &= 0.00102030304 \times \frac{1}{11} = 0.0000927 \text{ V} \quad (18) \end{aligned}$$

$$\begin{aligned} V_{5B}^+ &= -2 \frac{Z_o}{Z_o + Z_S} \\ &= -0.9090909 \text{ V} \quad (19) \end{aligned}$$

$$\begin{aligned} V_{in} &= V_1^+ + V_2^- + V_3^+ + V_4^- + V_{5A}^+ + V_{5B}^+ \\ &= 0.9090909 - 0.1010101 - 0.009182736 + 0.001020304 \\ &\quad + 0.0000927 - 0.9090909 \\ &= -0.10908 \text{ V} \quad 2 \text{ ns} < t < 3 \text{ ns} \quad (20) \end{aligned}$$

$$\begin{aligned} V_6^- &= (V_{5A}^+ + V_{5B}^+) \Gamma_L \\ &= (0.0000927 - 0.9090909) \times \frac{-1}{9} = 0.1009998 \text{ V} \quad (21) \end{aligned}$$

$$\begin{aligned} V_L &= V_1^+ + V_2^- + V_3^+ + V_4^- + V_{5A}^+ + V_{5B}^+ + V_6^- \\ &= 0.9090909 - 0.1010101 - 0.009182736 + 0.001020304 \\ &\quad + 0.0000927 - 0.9090909 + 0.1009998 \\ &= -0.00808 \text{ V} \quad 2.5 \text{ ns} < t < 3.5 \text{ ns} \quad (22) \end{aligned}$$

$$R = \frac{1}{2\pi} \sqrt{\frac{2\pi f \mu_o}{2\sigma_c}} \left(\frac{1}{b} + \frac{1}{a} \right) \Omega / m$$

$$L = \frac{\mu_o}{2\pi} \ln \left(\frac{b}{a} \right) H / m$$

$$G = \frac{2\pi}{\ln \left(\frac{b}{a} \right)} (2\pi f \epsilon_r \epsilon_o \tan \delta) \frac{1}{\Omega} / m$$

$$C = \frac{2\pi \epsilon_r \epsilon_o}{\ln \left(\frac{b}{a} \right)} F / m \quad (23)$$

$$Z_o = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \Omega$$

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} / m \quad (24)$$

the pulse edges as a function of cable loss.

For the most accurate measurements when evaluating a digital pulse, it is recommended to first determine the frequency components of the pulse, and then determine the bandwidth in which the pulse is primarily contained. Provided that the cable loss at the high end of the pulse bandwidth is not too high, it

can be assumed the coaxial cable will not significantly affect the pulse.

When the load appearing across the output of a circuit under test is of interest, it is possible to use the equations above to incorporate the parameters of a test cable as well as the input impedance of the test equipment. Mainly, the cable parameters are determined from Eqs. 23, 24, and 25. The impedance of the test

$$\begin{aligned} Cbl_{ABCD} &= \begin{pmatrix} Cbl_A & Cbl_B \\ Cbl_C & Cbl_D \end{pmatrix} \\ &= \begin{bmatrix} \cosh(\gamma_1 \ell_1) & Z_{o1} \sinh(\gamma_1 \ell_1) \\ \frac{1}{Z_{o1}} \sinh(\gamma_1 \ell_1) & \cosh(\gamma_1 \ell_1) \end{bmatrix} \\ &\quad \times \begin{bmatrix} \cosh(\gamma_2 \ell_2) & Z_{o2} \sinh(\gamma_2 \ell_2) \\ \frac{1}{Z_{o2}} \sinh(\gamma_2 \ell_2) & \cosh(\gamma_2 \ell_2) \end{bmatrix} \\ &\quad \times \begin{bmatrix} \cosh(\gamma_3 \ell_3) & Z_{o3} \sinh(\gamma_3 \ell_3) \\ \frac{1}{Z_{o3}} \sinh(\gamma_3 \ell_3) & \cosh(\gamma_3 \ell_3) \end{bmatrix} \quad (25) \end{aligned}$$

$$\begin{aligned} S_{11} &= \frac{Cbl_A + Cbl_B / Z_o - Z_o \times Cbl_C - Cbl_D}{Cbl_A + Cbl_B / Z_o + Z_o \times Cbl_C + Cbl_D} \\ SWR &= \frac{1 + |S_{11}|}{1 - |S_{11}|} \quad (26) \end{aligned}$$

$$S_{21} = \frac{2}{Cbl_A + Cbl_B / Z_o + Z_o \times Cbl_C + Cbl_D} \quad (27)$$

$$\Delta f \approx \frac{3 \times 10^8}{2\sqrt{2}} \times \frac{39}{36} = 115 \text{ MHz} \quad (28)$$

$$t_r^{in} = 80 \text{ ps}$$

$$t_r^{out} = 84 \text{ ps}$$

$$\begin{aligned} t_r^{cable} &\leq \sqrt{t_r^{out^2} - t_r^{in^2}} \\ &\leq 25.6 \text{ ps} \quad (29) \end{aligned}$$

$$t_d \approx \frac{\sqrt{2}}{3 \times 10^8} \times \frac{36}{39} = 4.4 \text{ ns} \quad (30)$$

$$C = \frac{2\pi \epsilon_r \epsilon_o}{\ln(b/a)} \times 36 \times \frac{2.54}{100} = 86 \text{ pF} \quad (31)$$

$$L = \frac{\mu_o}{2\pi} \ln(b/a) \times 36 \times \frac{2.54}{100} = 216 \text{ nH} \quad (32)$$

equipment, and subsequent load impedance to the circuit, can be determined from Eqs. 4, 5, and 6. **MRF**

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Specifying RF power amps for immunity and susceptibility testing

HIGH-POWER AMPLIFIERS ARE essential components in immunity and susceptibility test systems. While most specifiers are familiar with the standard characteristics for defining an amplifier's performance, such as gain, output power, and gain flatness, a less known characteristic—load tolerance—can play a major role in determining the ultimate performance from a power amplifier in an immunity/susceptibility test applications. An application note from Amplifier Research (Souderton, PA), "The importance of load tolerance in specifying RF power amplifiers for immunity/susceptibility testing applications" provides invaluable guidance in this area.

The application note introduces the concept of minimum available power to help specifiers understand how much power can be expected from a particular amplifier model even under various conditions of load mismatch. As the note details, the load shown to a power amplifier (PA) can vary widely over frequency due to variations in antenna characteristics, room reflections and resonances, imperfect connectors and cables, and reflec-

tions from a device under test (DUT). A test system with an antenna VSWR of 2.50:1 can easily contribute to a load VSWR in excess of 5.0:1 when other factors are considered.

In offering a solution to load mismatch problems, the note explains the differences between the types of amplifiers used for immunity/susceptibility testing: Class A and Class AB amplifiers. Class A amplifiers are designed to be load intolerant in these test applications. Although generally larger and more expensive than Class AB amplifiers, Class A designs provide superior electrical performance over a wider range of load mismatch conditions. For example, a Class A amplifier can absorb much higher levels of reflected power. Class A amplifiers are biased for extreme low distortion and excellent linearity.

The six-page application note offers a simple equation for calculating minimum available power, along with a variety of supporting performance plots. Copies are free for downloading from the company's website.

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Although generally larger and more expensive than Class AB amplifiers, Class A amplifiers provide superior electrical performance over a wider range of load mismatch conditions.

Characterize miniature baluns for wireless applications

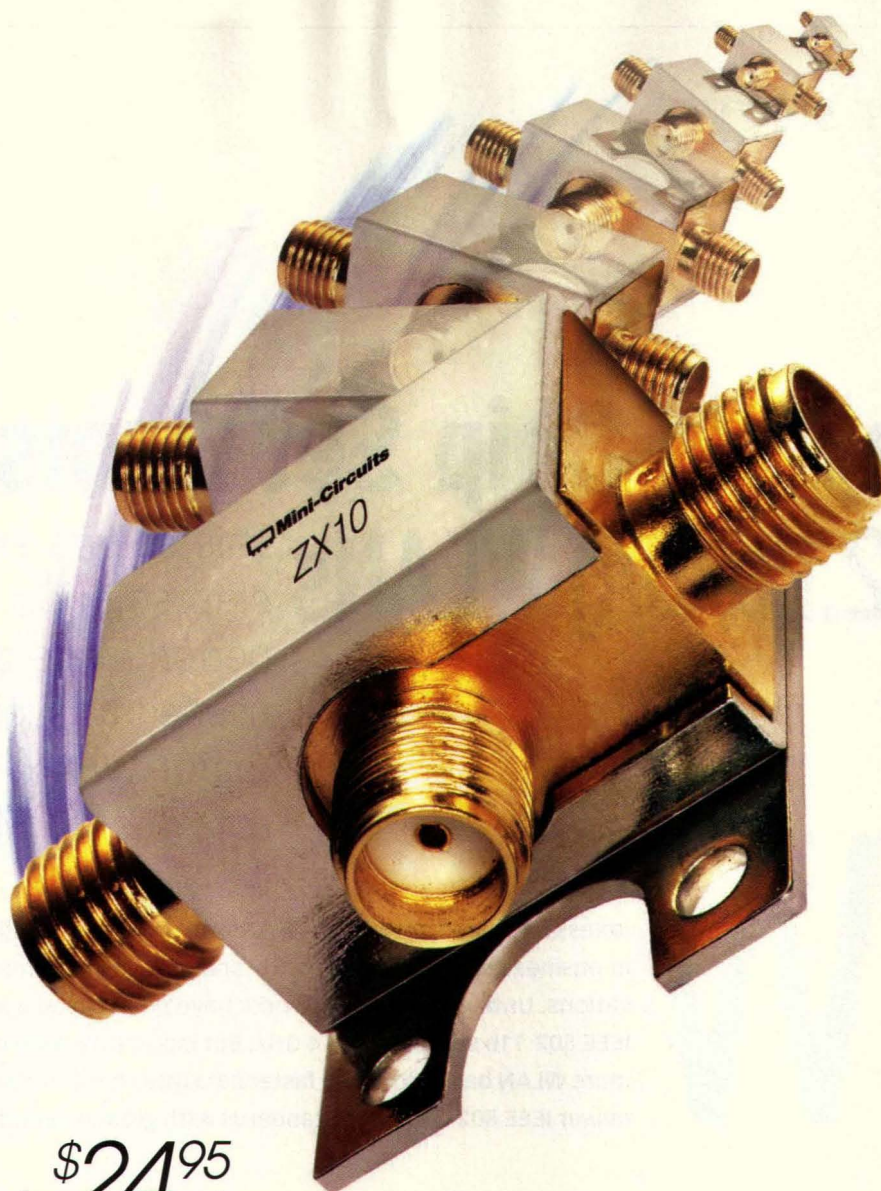
MINIATURE BALUNS PROVIDE THE critical impedance transformation between balanced and unbalanced components in a variety of circuits. Mixers are just one of the examples of a component that benefits from the use of low-loss miniature baluns. For those wishing to know more about the use of drop-in baluns in their designs, Anaren Microwave (East Syracuse, NY) offers a seven-page application note, "Mini balun transformer testing and characterization for commercial & consumer wireless application."

The miniature baluns are supplied in surface-mount packages that are designed to provide superior electrical performance to ceramic and lumped-element baluns. The wideband balanced-to-unbalanced transformers provide an unbalanced (single-ended) port impedance of 50 Ω and a balanced (differential) port impedance of 25 Ω to ground. The balanced outputs evenly distribute power from the unbalanced port providing a 180-deg. phase difference between the two output ports.

The application note defines and describes a variety of characteristics for these miniature baluns, including port impedance, return loss, isolation, insertion loss, and amplitude and phase balance. It details three characterization solutions for evaluating the baluns: the mixed-mode network-analyzer approach, the two-port network-analyzer approach, and the time-domain measurement method.

Additional Internet links are provided (such as to www.agilent.com) for those cases where more detailed information is needed in specifying test equipment or setting up the measurement system. In some of the approaches (such as the two-port measurement method), special test fixtures may be needed in order to measure amplitude and phase balance. Copies of the seven-page application note are available for free download from the company's website. For more information, contact:

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Flexible Chip Set Arms 802.11a/b/g WLANs

This software-driven, dual-chip solution provides the performance and versatility needed to support the three major WLAN standards at data rates to 54 Mb/s.



Wireless local-area networks (WLANs) are now well established in business enterprise networks, and growing in home applications. Until now, such networks have been based upon the IEEE 802.11b standard at 2.4 GHz. But increasing demands for more WLAN bandwidth and faster data rates have fostered the newer IEEE 802.11a and g standards with greater security and

better support of multimedia services. To enable these emerging WLAN standards, RF Micro Devices (Greensboro, NC) has developed the model RFCS5420 software-driven, flexible dual-band, two-chip solution that works across all three standards and allows seamless connectivity among all three systems.

A chip set designed for connectivity to IEEE 802.11a/b/g networks will require three main building blocks:

1. A radio subsystem with transceivers capable of operating at 2.4 and 5 GHz.
2. A modem that supports orthogonal-frequency-division-multiplex (OFDM) and complementary-code-keying (CCK) modulation schemes.
3. A unified media-access controller (MAC) with support for IEEE 802.11a/b/g and the extensions to these standards.

The RFCS5420 chip set consists of a model RF5425 transceiver integrated circuit (IC) and a model RF5421 bandband/MAC IC. Both ICs are implemented in 0.18- μm , +1.8-VDC complementary-metal-oxide-semiconductor (CMOS) technology which offers extremely low power consumption in both transmit and receive modes. The RFCS5420 chip set can be functionally viewed as a composite of distinct IEEE 802.11a and IEEE 802.11b/g chip sets, each with its own radio, modem, and MAC subsystems that share a common host interface. The RF5425 transceiver is based on a *True Zero-IF* CMOS radio-transceiver architecture. Only front-end amplifiers and bandpass filters are needed for a complete dual-band 2.4- and 5-GHz WLAN radio solution.

JIM BOHAC

Marketing Manager

RF Micro Devices (San Jose Office), 7628 Thorndike Rd., Greensboro, NC 27409; (336) 664-1233, FAX: (336) 931-7454, Internet: www.rfmd.com.

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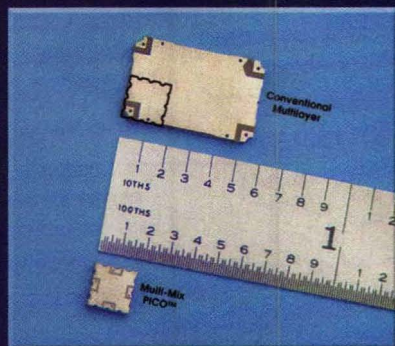
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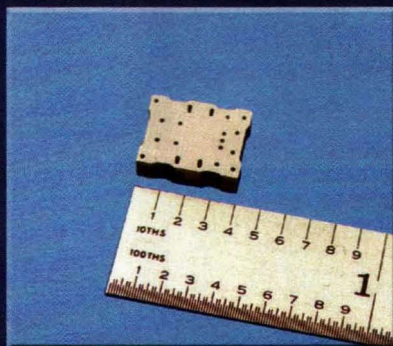
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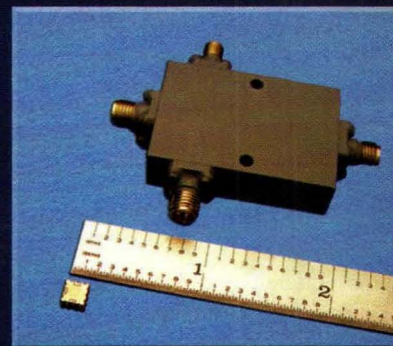
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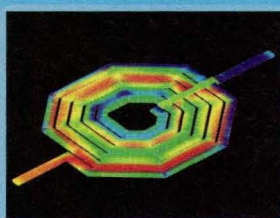
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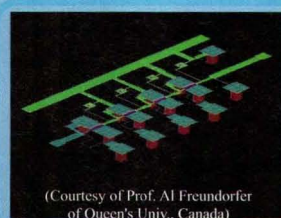
IE3D

PRODUCTS

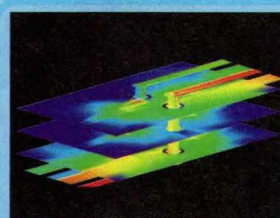
Planar and 3D Electromagnetic Simulation and Optimization Package



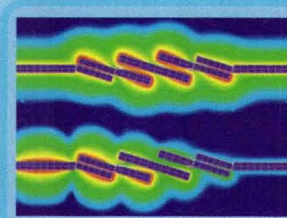
A Spiral Inductor with Thickness on a Semiconductor Substrate



(Courtesy of Prof. Al Freundorfer of Queen's Univ., Canada)



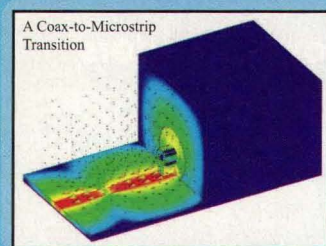
A Multilayer PCB with Vias in High-Speed Circuits



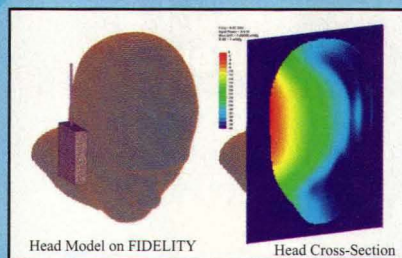
Near Field Visualization on IE3D 10.0

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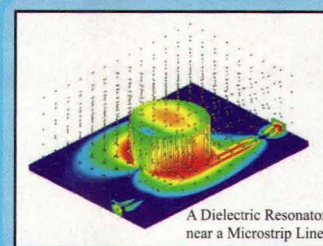


A Coax-to-Microstrip Transition



Head Model on FIDELITY

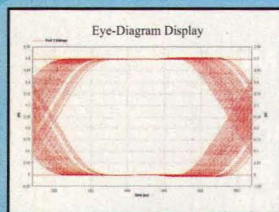
Head Cross-Section



A Dielectric Resonator near a Microstrip Line

MDSPICE

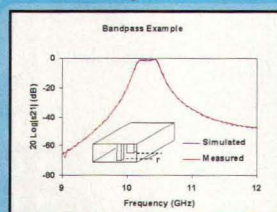
Mixed Frequency- and Time-Domain SPICE Simulator



Eye-Diagram Display

COCAFIL

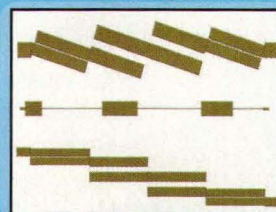
Waveguide Coupled Cavity Filter Design Suite



Bandpass Example

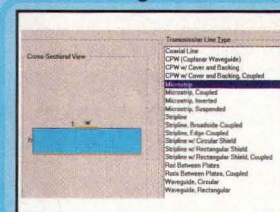
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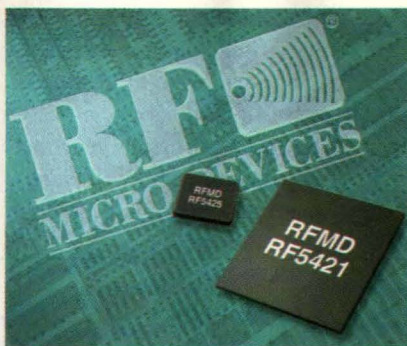
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The RF5421 baseband/MAC IC includes a complete implementation of IEEE 802.11a/b/g CCK and OFDM modems and an ARM9 processor that executes the bulk of the processing functions required for MAC processing. The RF5421 also includes hardware accelerators for modem aiding functions and full-speed encryption and security support. Together, the two ICs execute all of the Layer 1 and 2 functions required for IEEE 802.11a/b/g operation, thereby freeing significant high-speed processing chores from a host processor.

The RFC5420 chip set (Fig. 1) supports data rates to 11 Mb/s (1, 2, 5.5, and 11 Mb/s) in CCK mode at 2.4 GHz and data rates to 54 Mb/s (6, 9, 12, 18, 24, 36, and 54 Mb/s) in OFDM mode at 2.4 and 5 GHz. The radio transceiver operates in the 2.4-GHz industrial-scientific-medical (ISM) band, the 4.9-to-5.1 GHz Japan band, and



1. The RFC5420 WLAN chip set consists of the RF5425 2.4/5-GHz RF transceiver IC and the RF5421 baseband/MAC IC.

the 5.15-to-5.35-GHz UNII band. The chip set provides support for a wide range of modulation formats, including BPSK, QPSK, 16QAM, and 64 QAM. The chip set's unique AccuChannel equalization provides as much as 4-dB improvement in signal-to-noise ratio (SNR) in typical office environments,

which translates into a 32-percent increase in WLAN range and about 70-percent more coverage area when operating in OFDM mode.

The RF5425's zero-IF radio IC features a direct-conversion architecture (Fig. 2), requiring only one mixer stage to convert the desired RF signals directly to and from baseband signals without any IF stages and without the need for external surface-acoustic-wave (SAW) filters. Most zero-IF radio designs also integrate the low-noise amplifier (LNA), voltage-controlled oscillator (VCO), and the baseband filters on a monolithic die. In fact, such integrated single-chip zero-IF transceivers have performed well for many years in cellular and pager applications and they are beginning to emerge in WLAN radio designs as well. A major advantage of the zero-IF architecture on CMOS is that it enables the implementation of a full dual-band transceiver on a monolithic die with a

"Inject noise?, I'm trying to get rid of it!"



- **Noise figure systems for automated test equipment (ATE)**



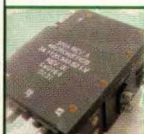
- **Calibrated noise sources for benchtop noise figure testing**



- **Built-in BER vs E_b/N_0 testing**



- **Dithering for A/D converters**



- **Built-in radar receiver test**

"Selling noise devices often elicits the following question, 'Why noise, I'm trying to get rid of it.' A good physical analogy is thinking of stars in space or light through stained glass. Stars can only be seen at night because the background light (noise) is too strong to see the small signal of the star; we see colors when we shine white light through stained glass. This is similar to injecting white noise through a filter and into a spectrum analyzer. Perhaps the most common noise application is simply using a calibrated noise source in conjunction with a bench-top test instrument to measure noise figure of an LNA, mixer or receiver front end. There are many other applications, call me to discuss yours."

*Pat Robbins, Director of Noise Products
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minimum increase in die size compared to a single-band implementation. The zero-IF radio architecture eliminates an IF stages, reducing complexity and power consumption. Furthermore, several on-chip circuits, i.e. synthesizers, can be shared for both bands.

Of course, no radio architecture is ideal. Some of the common RF problems inherent with the zero-IF architecture are DC offset, flicker noise, and LO pulling. DC offsets are mainly generated by the LO leakage, which self-mixes, thereby creating a DC component in the

signal chain that affects the receiver performance and can cause the RF stages to saturate. Flicker noise, also known as $1/f$ noise, is low-frequency device noise that can corrupt signals in the receiver chain. Flicker noise is more pronounced with the zero-IF architecture because of the direct conversion to low-frequency baseband signals. Another concern with direct conversion is the pulling of the LO by the power-amplifier (PA) output, which affects the direct upconversion process. This is because the high-power PA output, which has a spectrum centered around the LO frequency, can disturb (or pull) the frequency of the transmitter VCO.

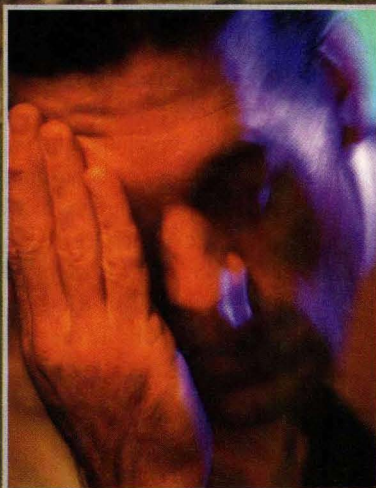
The RF5425 incorporates proprietary filters designed to provide superior adjacent-channel and alternate adjacent-channel rejection while minimizing noise contributions. An innovative VCO-design and frequency-planning architecture minimizes phase noise and LO pulling in the transmitter.

The other RF effects of zero-IF CMOS transceivers are mitigated by a combination of RF design techniques and baseband algorithms. A DC offset loop operates in conjunction with the baseband to dynamically correct for DC offset. Similarly, in-phase/quadrature (I/Q) mismatch is measured and corrected by the baseband at time of system initialization or channel changes. Close coupling of the RF5421 and RF5425 provide optimal performance for the zero-IF WLAN radio implementation.

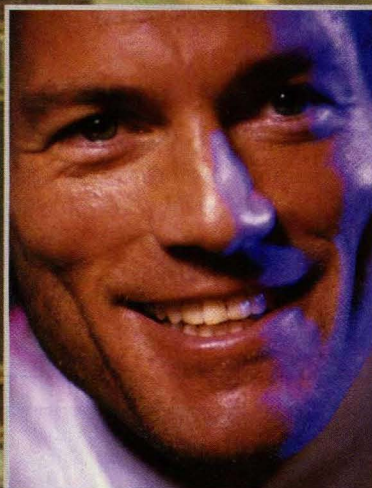
The RF5421 baseband/MAC IC provides the baseband, MAC, security and host functions for the RF5425 radio (Fig. 3). It is designed with proprietary RF Micro Devices algorithms to achieve maximum IEEE 802.11a/b/g system performance, using AccuChannel Equalization technology, advanced hardware security and a flexible MAC. The IC incorporates the following functional blocks:

1. An ARM9 processor with 16-kB cache memory for MAC software execution supported by external SDRAM, SRAM, or Flash memory.
2. An 802.11a/g OFDM modem.
3. An 802.11b CCK modem.

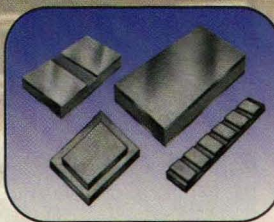
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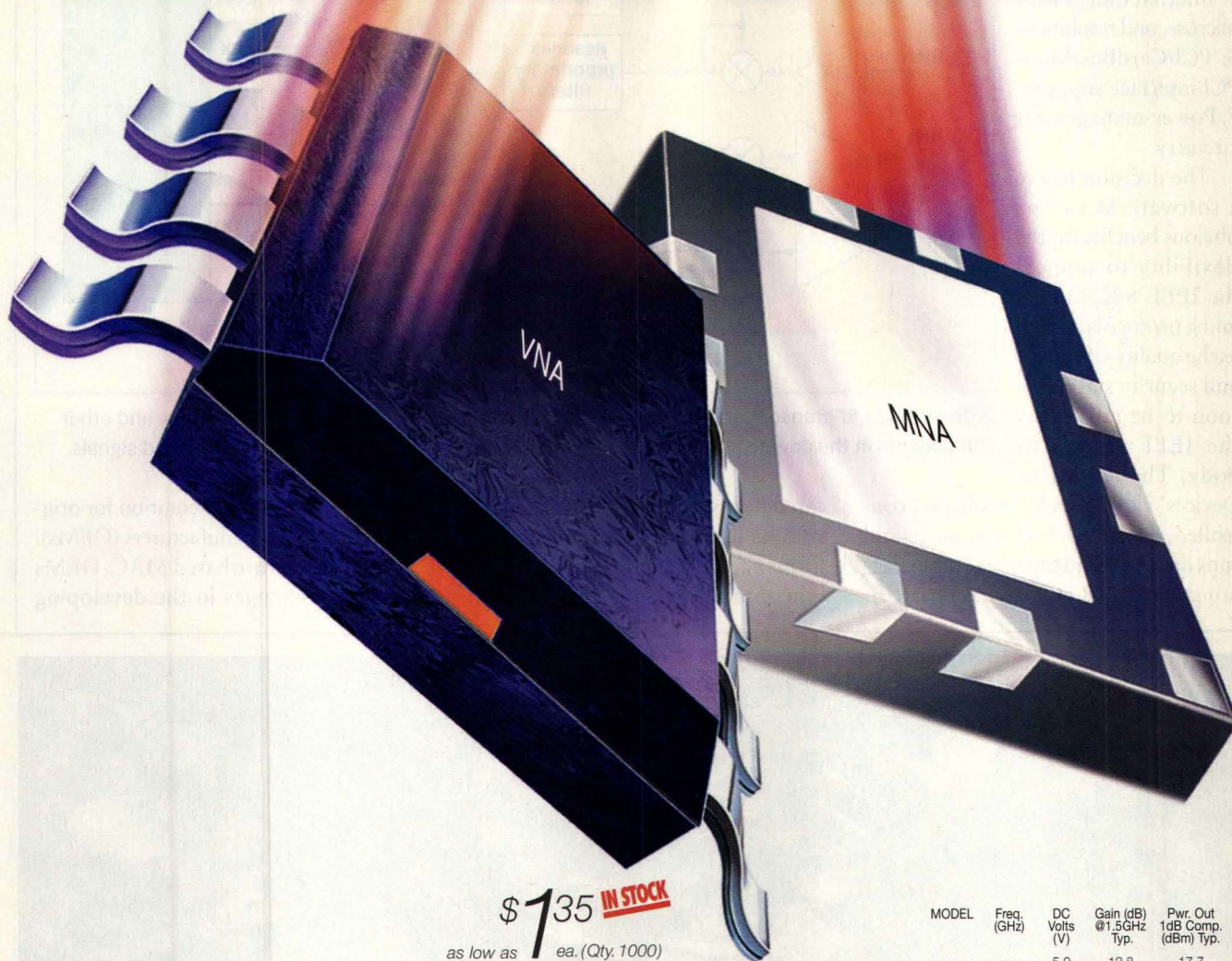
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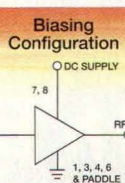
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	MNA-4	0.5-2.5	5.0 2.8	16.4 14.5	19.0 13.4	1.90
	MNA-5	0.5-2.5	5.0 2.8	21.9 20.5	12.2 10.1	1.60
	MNA-6	0.5-2.5	5.0 2.8	23.6 21.2	18.0 14.1	2.25
	MNA-7	1.5-5.9	5.0 2.8	15.9 13.7	15.6 12.7	2.25
	VNA-21	0.5-2.5	5.0 2.8	13.5 12.3	8.5 7.0	1.80
	VNA-22	0.5-2.5	5.0 2.8	13.8 12.6	17.0 14.0	2.20
	VNA-23	0.5-2.5	5.0 2.8	18.3 17.1	10.0 8.5	1.90
	VNA-25	0.5-2.5	5.0 2.8	18.6 17.4	18.2 12.0	2.50
	VNA-28	0.5-2.5	5.0 2.8	22.8 21.0	11.0 9.6	1.95

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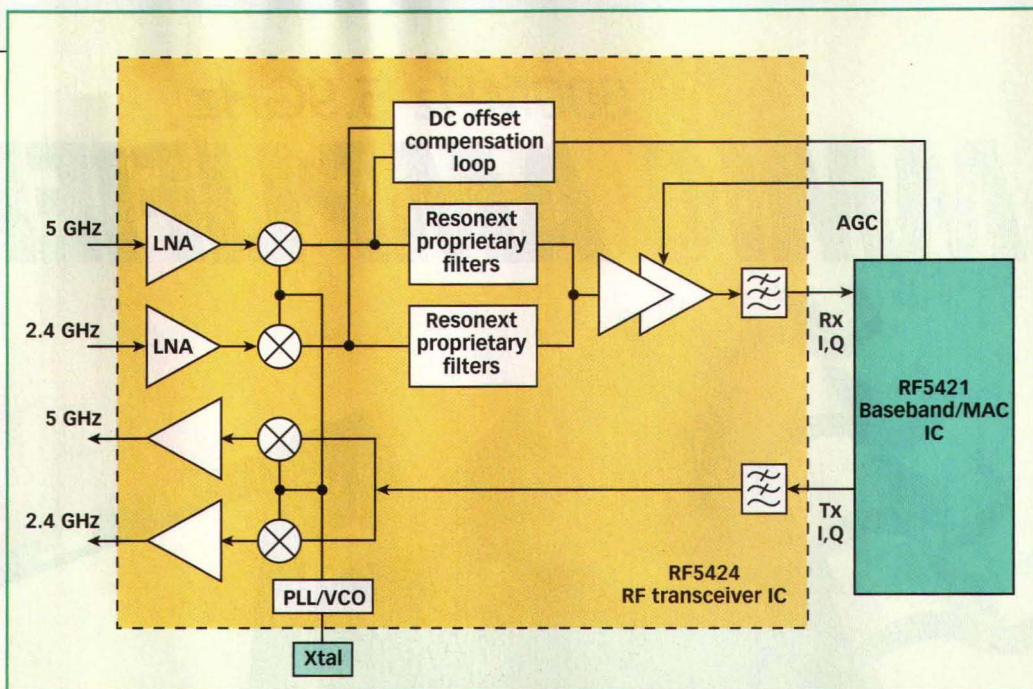
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The decision to use a software MAC has obvious benefits for the flexibility to support the IEEE 802.11a, b, and g protocols as well as the quality-of-service and security standards soon to be ratified by the IEEE standards body. The RF Micro Devices' Flexible MAC is software controlled, written in C++ language, and runs under the ThreadX real-time operating system (RTOS). On-chip dedi-

cated hardware is provided to offload various low-level operations for performance optimization. Implementation of the MAC in software provides

a *future-proof* implementation for original-equipment manufacturers (OEMs). With the software based MAC, OEMs can track changes in the developing



2. The RF5425 RF transceiver IC employs a zero-IF architecture to minimize mixers, filters, and other components in the downconversion of 2.4- and 5-GHz signals to lower-frequency baseband signals.

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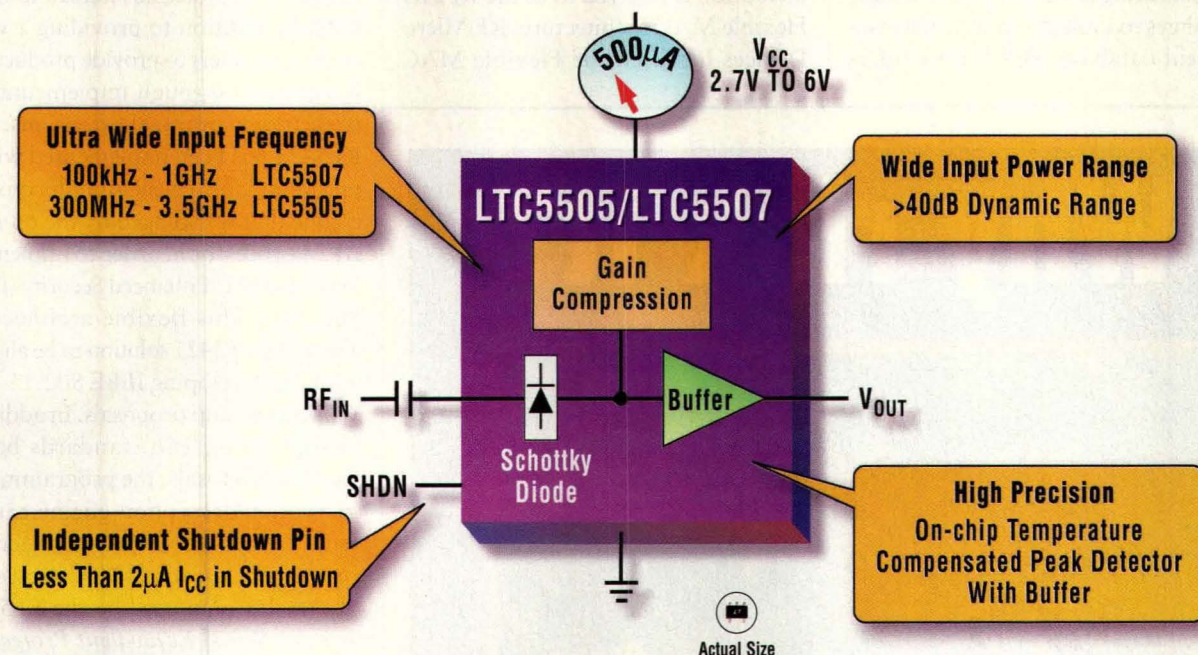
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Infrastructure and Handset

LT5504	800MHz to 2.7GHz RF Measuring Receiver with 80dB Dynamic Range
LT5511	High Signal Level Upconverting Mixer with +17dBm Input IP3
LT5512	High Signal Level Downconverting Mixer with +17dBm Input IP3
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IEEE 802.11 standards and provide feature upgrades to installed hardware via software-update packages. All flexible MAC features can be downloaded to clients using remote software-update procedures controlled from a central management database. This feature offers

OEMs the capability to provide software updates and revenue-generating software enhancements to end customers.

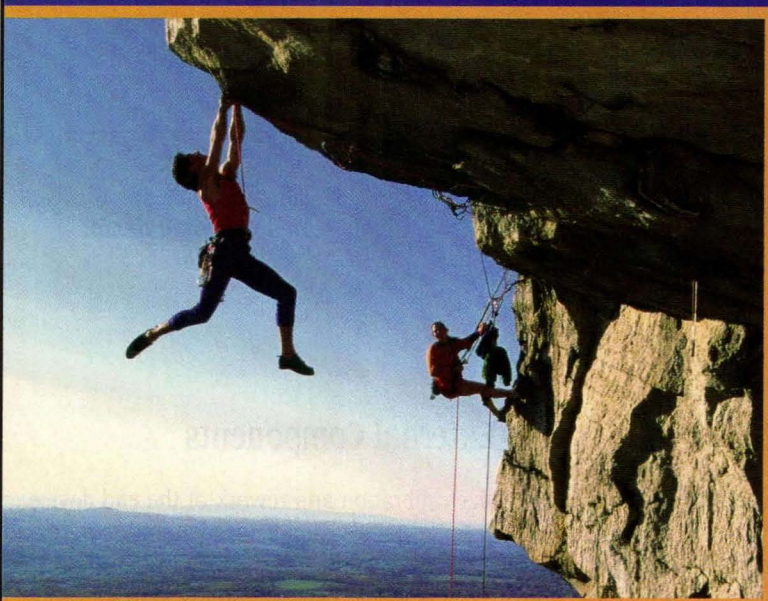
This software-based MAC implementation is referred to as the RFMD Flexible MAC architecture. RF Micro Devices has used the Flexible MAC

architecture to track IEEE standard developments (e.g., IEEE 802.11i,e,h,k, and j) and to implement extensions to the basic service set defined by the IEEE such as WiFi Protected Access and Cisco CCX in addition to providing a vehicle for customers to provide product differentiation through implementation of their own protocol extensions. The RF5421 has been implemented with a programmable encryption/decryption engine that is managed by the on-chip ARM processor in order to implement Flexible MAC Enhanced Security (IEEE 802.11i). This flexible architecture allows the RF5421 solution to be aligned with the developing IEEE 802.11-TGi Working Group proposals. In addition to supporting TGi standards based security proposals, the programmable encryption/decryption engine can be used to support proprietary security systems such as Cisco CCX. Current security modes supported by the RF5421 include: *Wired Equivalent Protection* (WEP) 1 and 2, *WiFi Protected Access* (WPA) and *Advanced Encryption Standard* (AES). These security modes are implemented in hardware for maximum system performance. TKIP and AES/CCM modes are also supported for future compatibility with evolving TGi proposals. The RF5420 chip set also provides support for 802.1x authentication.

Multimedia applications are increasing the need for managed quality of service (QoS) in many applications. In accordance with the IEEE 802.11e standard, the RFMD Flexible MAC has provisioned for eight independently managed packet queues to support implementation of QoS dependent services. With the software-based architecture the Flexible MAC can be used to implement standards-based approaches to QoS or proprietary approaches in advance of full standard ratification.

The IEEE 802.11 specifications allow for reserved and optional fields in beacon and other management frames within the IEEE 802.11 protocols. OEMs have taken advantage of these fields to implement network-management features commonly referred to as "SuperMAC" extensions. RFMD's Flexible MAC

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	HMC220MS8	+10 LO, DBL- BAL	5.0 - 12.0	DC - 4.0	-7.5	23	+17	\$1.99
	HMC218MS8	+7 LO, DBL- BAL	4.5 - 6.0	DC - 1.6	-8	28	+13	\$1.43
	HMC264LM3	Low LO, Sub-Harmonic	20 - 30	DC - 4.0	-9	30	+10	CALL
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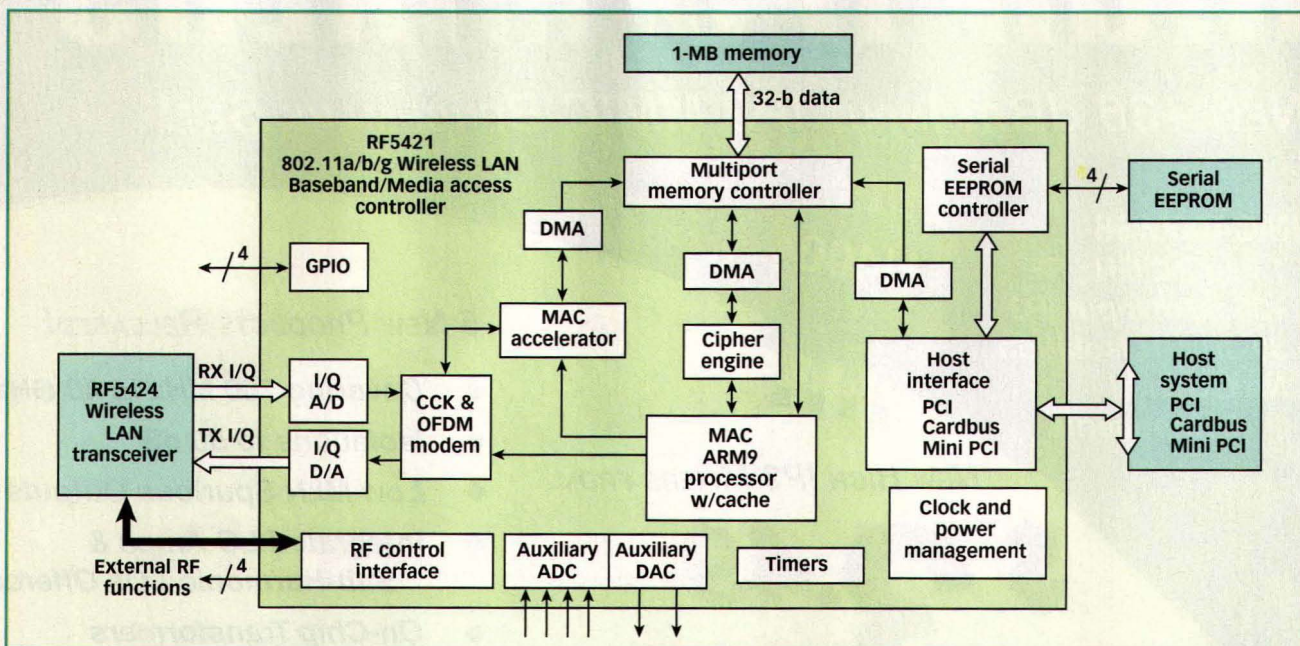
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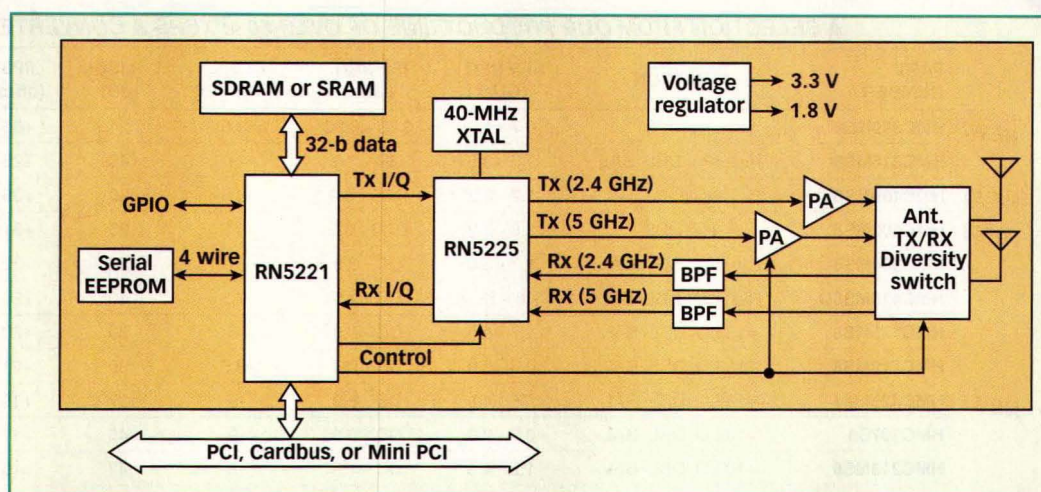
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3. The RF5421 baseband/media-access-controller (MAC) IC works with a flexible software MAC to support the IEEE 802.11a/b/g WLAN standards.

4. This functional block diagram shows a typical client CardBus or Mini-PCI WLAN design based on the RF5425 transceiver and the RF5421 baseband/MAC ICs.

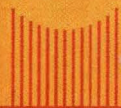


allows OEMs to take full advantage of these SuperMAC extensions. Examples of these features include dynamic data-rate selection, client roaming support, and a host of network-management features, such as link status reporting, and radio-link calibration.

With the Flexible MAC software and the two-chip CMOS implementation, the RFCS5420 chip set can be embedded into a variety of WLAN devices. As an example, Fig. 4 shows a functional block diagram for a basic client CardBus or Mini-PCI WLAN solution based on the RF5425 and RF5421. This application illustrates

the level of system integration achievable with the chip set resulting in a high-yield, low-BOM cost, small form-factor design. In addition to the pair of ICs, the solution requires dual-band diversity antennas, an antenna switch for transmit/receive mode and diversity functions, 2.4- and 5-GHz PAs, low-dropout regulators to supply the +1.8 and +3.3 VDC needed by all devices, a low-cost band-select filter/diplexer for selecting 2.4- or 5-GHz band, 2-kb serial EEPROM for storing device configuration data and MAC address, and 1-MB SDRAM or SRAM for MAC code and data.

In addition, the RFCS5420 chip set can be used to develop USB2.0 client devices, Ethernet Client Bridge devices, low-cost Access Points and Wireless routers, as well as embedded multimedia WLAN NICs. In summary, the RFCS5420 is an ideal solution to provide cost-effective IEEE 802.11a/b/g network connectivity while at the same time providing a platform that will change along with the continuously evolving IEEE standards activities. RF Micro Devices, 7628 Thorndike Rd. Greensboro, NC 27409; (336) 664-1233, FAX: (336) 931-7454, Internet: www.rfmd.com.



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This modular test set acts as a miniature radar system to make simple yet accurate amplitude and phase-pulse-stability measurements on two-port devices.

Radar component and system testing usually evokes images of complex, expensive measurement systems occupying multiple equipment racks. In some cases, radar equipment designers have even used entire radar systems as a proving ground for their new components and modules. With the introduction of the PN9002 pulse-to-pulse radar stability test set from Aeroflex, Inc. (Plainview, NY), radar testing

has become significantly less complex and costly. The modular PN9002 system provides outstanding dynamic range at frequencies from 2 to 18 GHz and optionally from 0.4 to 18 GHz.

The two-rack PN9002 replaces "homegrown" radar measurement systems often costing as much as \$1 million. Instead, the PN9002 is a fully self-contained, cost-effective system housed

in two compact VXI racks (see figure). The system delivers precise pulsed signals to a unit under test (UUT) and

then digitizes and analyzes the effect of the UUT on the test pulses. The PN9002 produces pulses at a maximum level of +13 dBm (with ± 2 -dB flatness), and requires UUT output levels from 0 to +2.5 dBm back at the input of the PN9002 for proper analysis.

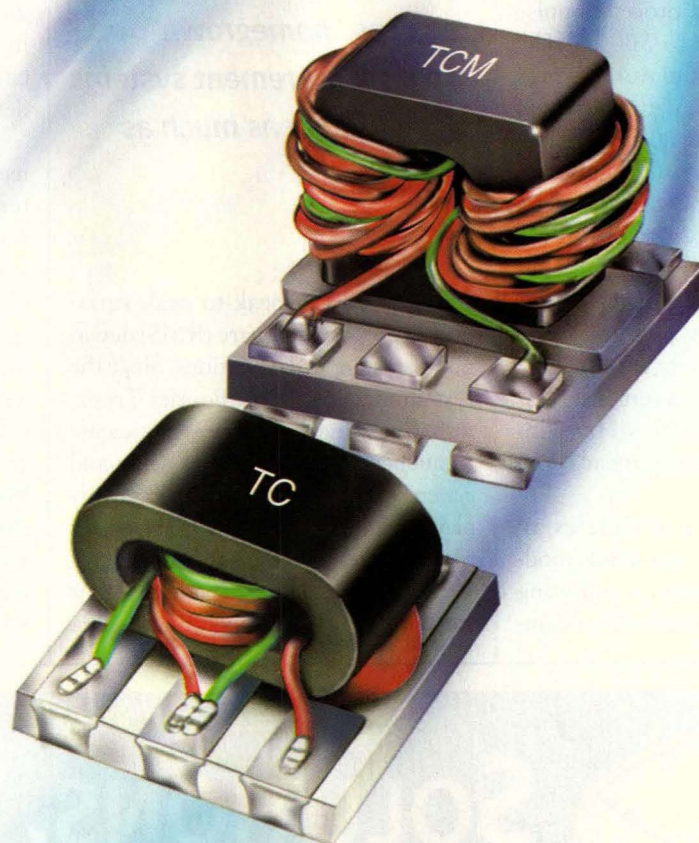
The PN9002 can analyze from 4 to 16 pulses during a measurement, using

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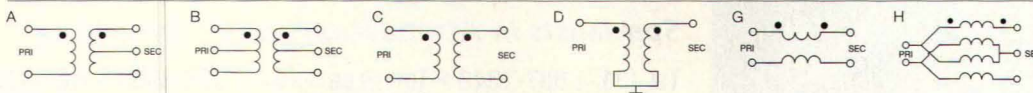
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TC3-1T	3A	5-300	5-300	1.29	
TC4-1T	4A	5-300	1.5-100	1.19	
TC4-1W	4A	3-800	10-100	1.19	
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TC8-1	8A	2-500	10-100	1.19	
TC9-1	9A	2-200	5-40	1.29	
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TCML1-19	1G	800-1900	900-1400	1.09	
TCM2-1T	2A	3-300	3-300	1.09	
TCM3-1T	3A	2-500	5-300	1.09	
TTCM4-4	4B	0.5-400	5-100	1.29	
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TCM4-6T	4A	1.5-600	3-350	1.19	
TCM4-14	4A	200-1400	800-1000	1.09	
TCM4-19	4H	10-1900	30-700	1.09	
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anywhere from 4 to 128 samples per pulse (depending upon the pulse width) to digitize the analog pulses (at sampling clock frequencies to 50 MHz). The test system generates and processes pulse widths from 200 ns to 500 μ s. The PN9002's pulse modulator features rise/fall time of 10 ns, with on time of 30 ns and on/off ratio of 80 dB. The integral pattern generator offers minimum pulse widths of 200 ns with 10-ns rise time at transistor-transistor-logic (TTL) output levels. The system achieves impressive residual noise levels of better than -76 dBc and typically -79 dBc with system-measurement accuracy of ± 2 dB. It covers a maximum instantaneous bandwidth of 20 MHz (with bandwidth adjustable by means of a digital Bessel filter).

The test system demodulates all phase-modulation and amplitude-modulation (AM) variations, computing key pulse-to-pulse or intrapulse param-

***The two-rack PN9002
replaces "homegrown"
radar measurement systems
often costing as much as
\$1 million.***

eters in terms of peak-to-peak variations, root-mean-square (RMS) deviations, or standard deviations. Since the system includes a Fast Fourier Transform (FFT) analyzer, it can also execute frequency-domain measurements and generate results in terms of spectral displays, working with user-defined Doppler filters.

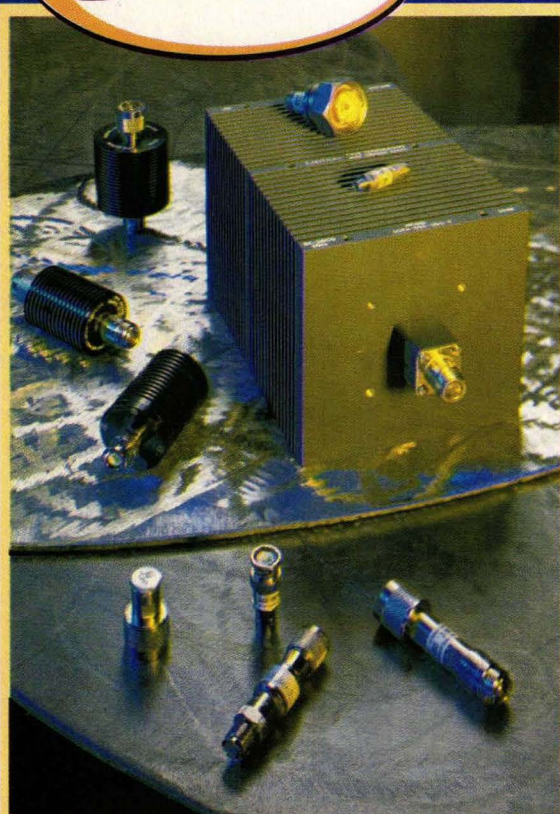
Software for the system is written for the Windows XP Pro operating system.

The software supports a variety of FFT windowing functions, including Hanning, Hamming, and Blackmann functions. The software also simplifies time-domain processing functions, including multiple interpolations, peak-to-peak calculations, RMS calculations, and determinations of standard deviations.

The basic ("starting") PN9002 system includes a generator rack with a pattern generator, low-noise frequency synthesizer, and pulse modulator, as well as an analyzer rack with a noise module, phase/amplitude-modulation detector, video shifter/low-noise amplifier (LNA), and signal-processing module. The basic system also includes a programmable phase shifter, control unit with display, mouse, and keyboard. P&A: \$160,000 and up; 12 wks. **Aeroflex, Inc., 35 South Service Rd., Plainview, NY 11803; (516) 694-6700, FAX: (516) 694-4823, Internet: www.aeroflex.com.**



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Smart Clocks Set Timing Standards

Each of these "intelligent" rubidium clocks integrates a wide range of timing and synchronization functions within a single compact housing.

Timing is everything in most telecommunications systems. Communications systems such as code-division-multiple-access (CDMA) cellular and synchronous-optical-network (SONET) systems rely on precise timing of transmitted data for proper operation. In a wireless environment, a lack of precise timing results in service-affecting dropped calls. In order to achieve that timing, these and other systems

auto-calibrate the frequency in case of power failure or loss of reference and to measure time-interval-error (TIE)

depend upon a precision atomic clock such as the iSync+™ Smart SRO-100 and SRO-75 rubidium (Rb) SynClock+™ standards from Temex Time. These smart synchronized-rubidium-oscillator (SROs) clocks can be seamlessly and auto-adaptively disciplined to a multi-vendor stratum 1 reference such as Global Positioning System (GPS), Cesium, LORAN-C, CDMA, and E1/T1 at industry's first 1-ns resolution thanks to its SmarTiming+™ technology, which

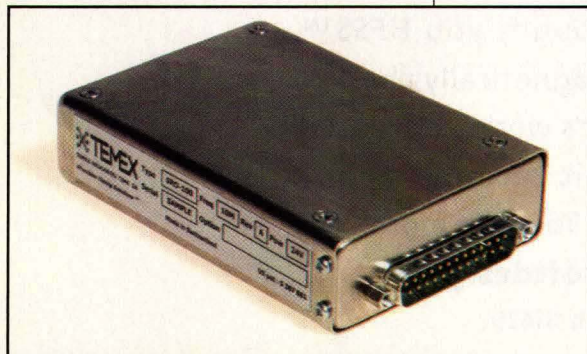
controls parameters such as GPS's Time-RAIM and position hold and filters input noise such as jitter and wander dynamically up to 100,000 s for optimal output performance. The technology also provides a sync or track mode to phase or frequency align the output to the refer-

ence and to adjust the output offsets up to 1 s with a 1-s comparator. The SRO has also an integrated EEPROM to

performance. In addition, it provides both sinewave and complementary-metal-oxide-semiconductor (CMOS)-level output signals with excellent short- and long-term stability, low current consumption, and fast warm-up times. The SRO-100 and SRO-75 are actually complete miniature synchronization systems, rather than simple Rb oscillators, generating CMOS, sinewave, and 1 pulse per second (PPS) signals as well as time-of-day information. In addition to the Rb electronics, they incorporate locking, disciplining and synchronization circuitries, bus control, EEPROM, and direct-digital-synthesizer (DDS) circuitry. The SRO-100 is available with sinewave and CMOS outputs in a housing with volume of 11 in.³ while the SRO-75 provides CMOS output signals in a volume of 5.5 in.³

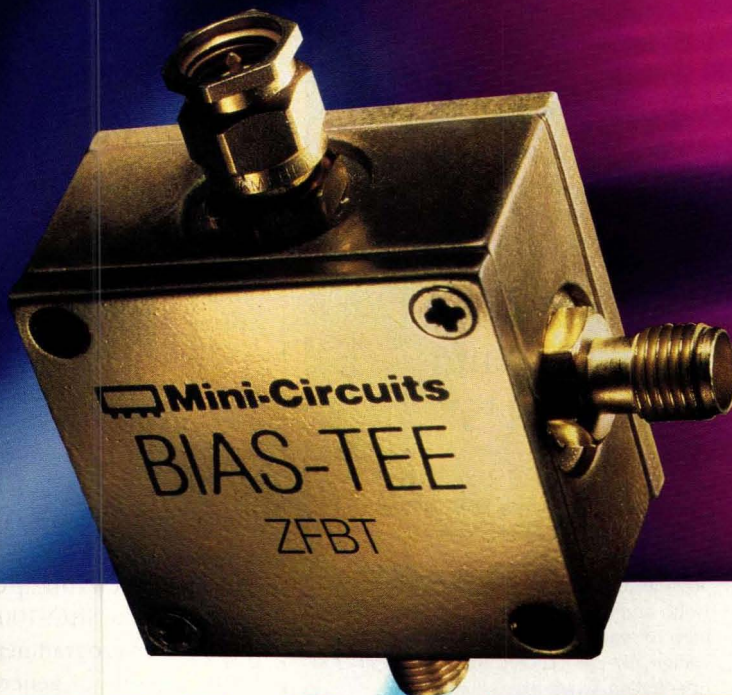
The clocks run in one of two modes: sync or track. In sync mode, an SRO-100 or SRO-75 phase aligns the output to the reference. In case of loss of reference, the holdover feature for the SRO-100 is less than 2 μ s over 48

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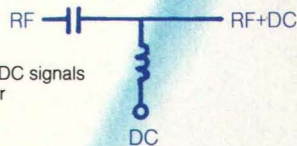
The SRO-100 is a self-contained Rb clock and synchronization system with standard 10-MHz and 1-PPS output signals.

ence and to adjust the output offsets up to 1 s with a 1-s comparator. The SRO has also an integrated EEPROM to



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▲ZFBT-6GW	0.1-6000	0.15	0.6	1.0	25	40	30	1.13:1	89.95
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▲ZFBT-6G-FT	10-6000	0.15	0.6	1.0	N/A	N/A	N/A	1.13:1	79.95
▲ZFBT-4R2GW-FT	0.1-4200	0.15	0.6	0.6	N/A	N/A	N/A	1.13:1	79.95
▲ZFBT-6GW-FT	0.1-6000	0.15	0.6	1.0	N/A	N/A	N/A	1.13:1	89.95
★ZNBT-60-1W	2.5-6000	0.2	0.6	1.6	75	45	35	1.35:1	82.95
■PBTC-1G	10-1000	0.15	0.3	0.3	27	33	30	1.10:1	25.95
■PBTC-3G	10-3000	0.15	0.3	1.0	27	30	35	1.60:1	35.95
■PBTC-1GW	0.1-1000	0.15	0.3	0.3	25	33	30	1.10:1	35.95
■PBTC-3GW	0.1-3000	0.15	0.3	1.0	25	30	35	1.60:1	46.95
■JEBT-4R2G	10-4200	0.15	0.6	0.6	32	40	40	-	39.95
■JEBT-6G	10-6000	0.15	0.7	1.3	32	40	40	-	59.95
■JEBT-4R2GW	0.1-4200	0.15	0.6	0.6	25	40	40	-	59.95
■JEBT-6GW	0.1-6000	0.15	0.7	1.3	25	40	30	-	69.95

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hours, and less than 7 μ s over 48 hours for the SRO-75. When locked to a stratum 1 reference, both clocks exceed the MTIE and TDEV masks defined by ITU-T G.811/823 and ANSI T1.101 standards. In track mode, an SRO-100 or SRO-75 frequency aligns the out-

put to the reference. In either mode, an SRO-100 or SRO-75 can adjust the time or phase offset of the output to up to 1 s through a 1-ns resolution comparator, operating over a dynamic range of ± 500 ns.

The SRO-100 (see figure) oper-

ates from a single voltage supply of +11 to +16 VDC or +18 to +32 VDC with standard sinewave output of 10 MHz (5 and 15 MHz are optional) and 60-MHz CMOS outputs. The frequency offset over temperature is $\pm 3 \times 10^{-11}$ at temperatures from -20 to $+60^\circ\text{C}$. The clock requires less than 1.2 A current during warm-up. The long-term stability is better than 5×10^{-11} /month and typically better than 3×10^{-11} /month. The short-term stability is 3×10^{-11} /s, 1×10^{-11} /10 s and 3×10^{-12} /100 s. The phase noise is -75 dBc/Hz offset 1 Hz from a 10-MHz carrier, -95 dBc/Hz offset 10 Hz from the same carrier, -125 dBc/Hz offset 100 Hz from the same carrier, -145 dBc/Hz offset 1 kHz from the same carrier, and -145 dBc/Hz offset 10 kHz from the same carrier. Harmonics are less than -25 dBc while spurious content is less than -80 dBc.

The SRO-75 provides 60-MHz CMOS output signals. It operates from a single voltage supply of +11 to +16 VDC or +18 to +26 VDC, and requires less than 0.8 A warm-up current at +24 VDC. Both the SRO-100 and SRO-75 feature user-programmable DDS circuitry capable of generating signals from DC to 20 MHz with 32-b digital resolution, as well as RS-232C ports for computer control.

The clocks are ideal for telecom synchronization applications in CDMA, synchronous-digital-hierarchy (SDH), and SONET, as well as tracking and guidance control, analog- or digital-television synchronization, military systems, and navigation systems. They are designed for reliable operation (the Rb lamp life expectancy is 20 years) and are equipped with an analog frequency-adjustment range (by means of a DC voltage of 0 to 5 V) of $5 \times 10^{-9} \pm 20$ percent. Jumpstart SRO-100 and Jumpstart SRO-75 Designer Kits are also available (with documentation and software) to help familiarize engineers with the operation of these clocks, while testing performance and validating their system design concept. Temex Time, Round Rock, TX; (512) 238-3127, FAX: (512) 238-3128, Internet: www.temextime.com.

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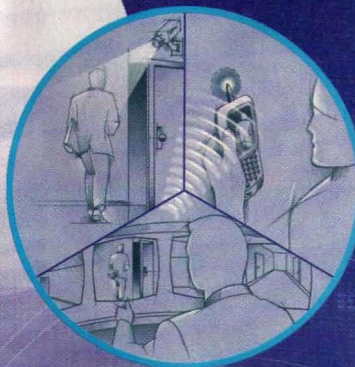
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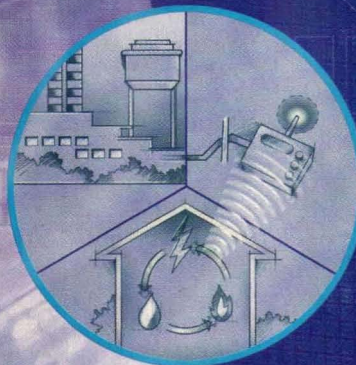
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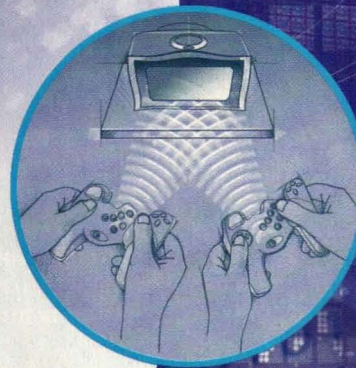
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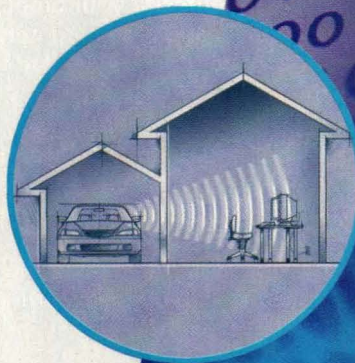
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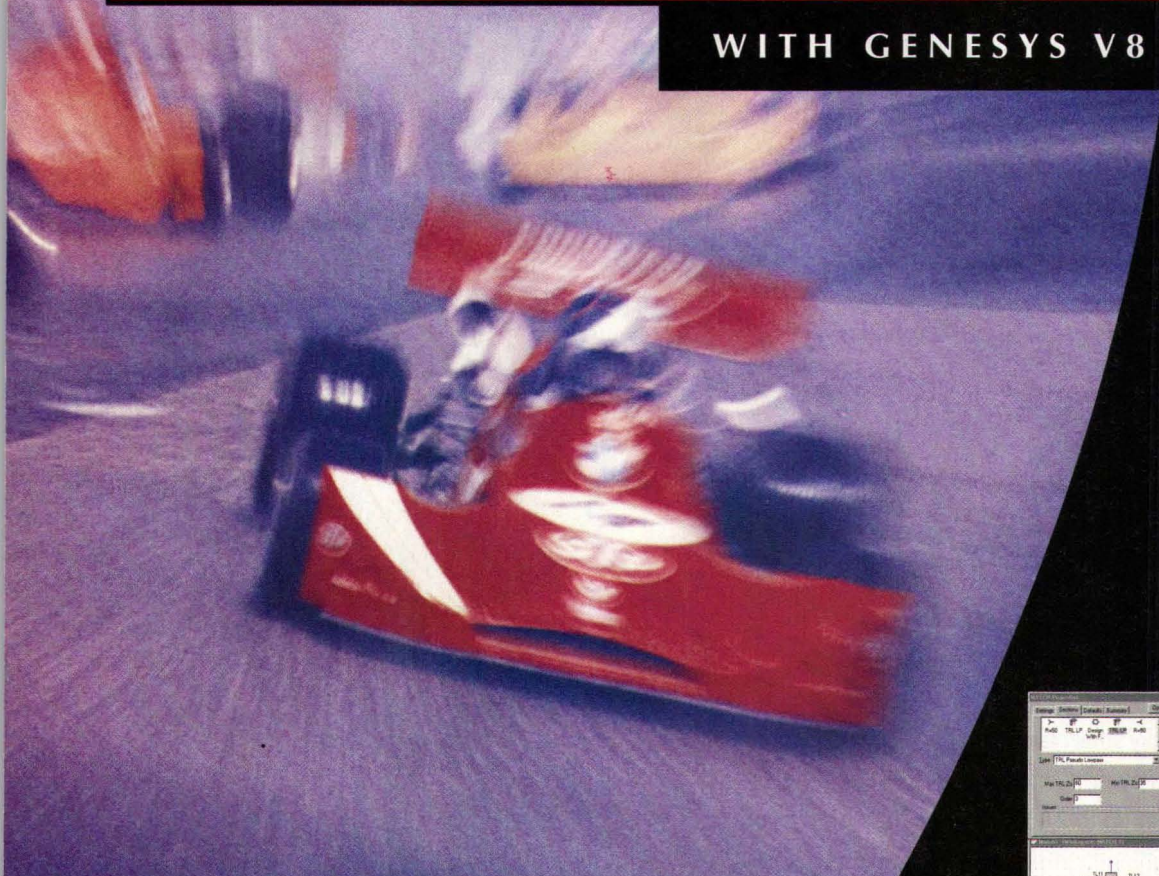
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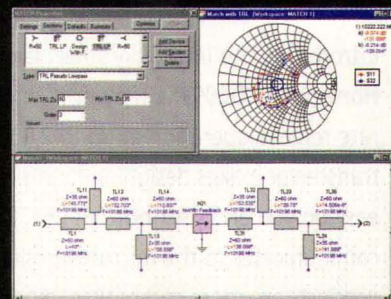
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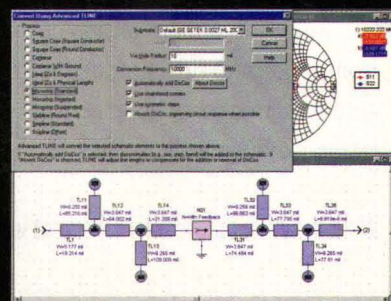
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Blind-Mate Connectors Bridge DC To 40 GHz

These high-frequency components provide reliable threadless connections with low insertion loss and ease of installation.

blind-mate connectors simplify RF signal interfaces by allowing push-on links between different components or elements of a design. A series of blind-mate connectors from MCE/Weinschel (Frederick, MD) goes a step beyond conventional blind-mate connectors by incorporating a unique spring-loaded mechanism to ensure reliable electrical contacts, even with some mechanical misalignment, at

frequencies from DC to 40 GHz.

The company's new blind-mate connectors (see figure) include the models 7008 (for pressurized applications), 7034 (for rear-locking installations), 7035 (for front-locking installations), and 7041 (a low-cost connector version for applications from DC to 18 GHz). The connectors are used in pairs, with one half of the pair providing a floating blind-mate interface with spring-loaded inner and outer contacts and the other half of the pair providing a fixed

blind-mate interface with fixed inner and outer contacts. These threadless connectors are designed to save

space, especially in multiple-connector, panel-mount applications such as receivers (Rx's), while providing excellent electrical performance and reliable operation (the blind-mate connectors are typically rated to withstand as many as 25,000 mating operations). Each connector pair can tolerate typically 0.02 in. of radial and axial offset misalignment per pair while still meeting minimum electrical specifications.

For applications requiring as much as 1000 psi hydrostatic pressure (and

JACK BROWNE
Publisher/Editor



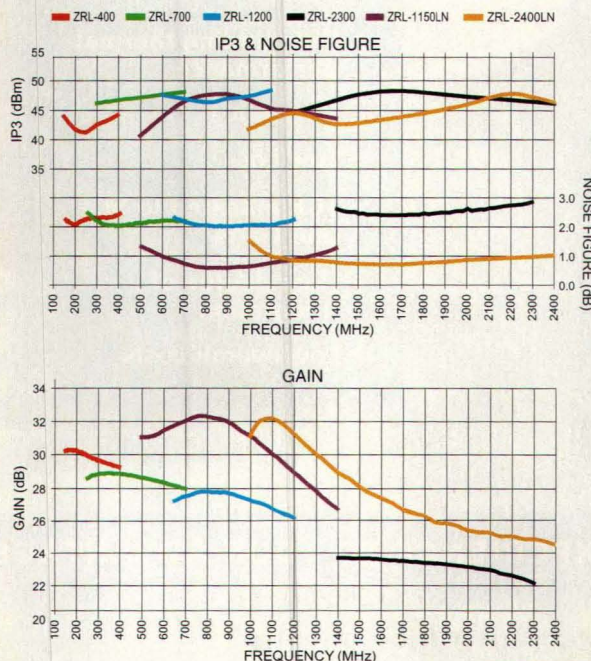
The MCE/Weinschel 40-GHz blind-mate connectors incorporate a spring-loaded mechanism to ensure reliable electrical contacts even with some mechanical misalignment.



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ZRL-1150LN	500-1400	31	0.8	45	24.0	139.95
ZRL-1200	650-1200	27	2.0	46	24.3	119.95
ZRL-2300	1400-2300	24	2.5	46	24.6	119.95
ZRL-2400LN	1000-2400	27	1.0	45	24.0	139.95

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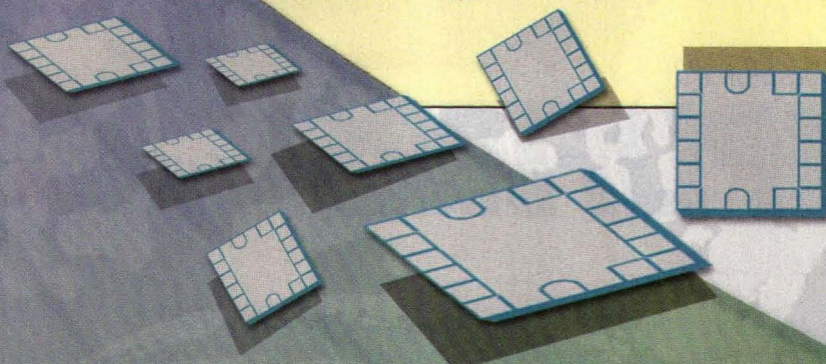
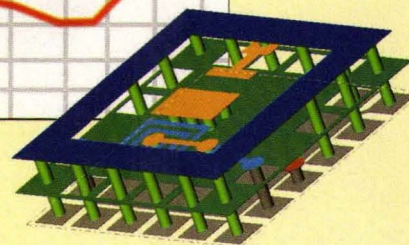
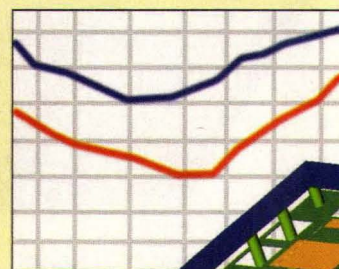
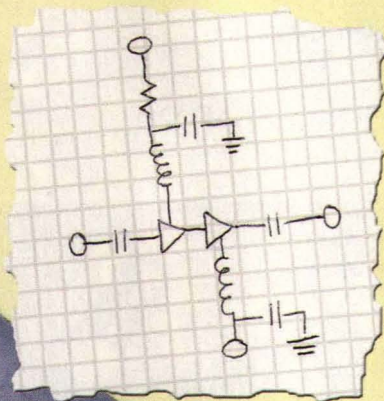
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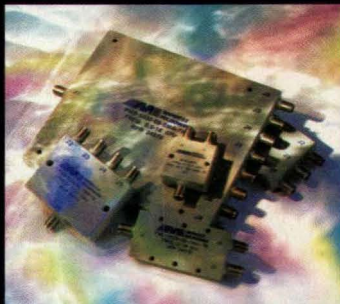
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50 psi static pressure), the model 7008 blind-mate connectors are 50-Ω SMA-type stainless-steel designs with an operating temperature range of -50 to +125°C. The connectors feature insertion loss of no greater than 0.3 dB through 18 GHz, no more than 0.8 dB

through 26.5 GHz, and no higher than 1.5 dB through 40 GHz. The VSWR is 1.30:1 or less through 18 GHz, 1.40:1 or less through 26.5 GHz, and 1.65:1 or less through 40 GHz. These pressurized planar blind-mate connectors are manufactured according to the interface

dimensions of MIL-STD-348 and are designed to mate nondestructively with MIL-C-39012 connectors.

For installations in which connectors must be locked by means of rear-panel access, the rear-locking models 7034 and 7034-1 provide maximum insertion loss of 0.85 dB through 40 GHz with less than 1.35:1 VSWR through 18 GHz and less than 1.55:1 VSWR through 40 GHz. Offering similar electrical performance, the models 7035, 7035-1, and 7035R-1 are designed for front-locking installations. Both connector

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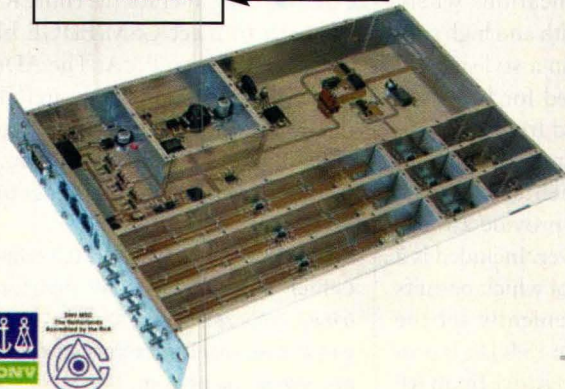
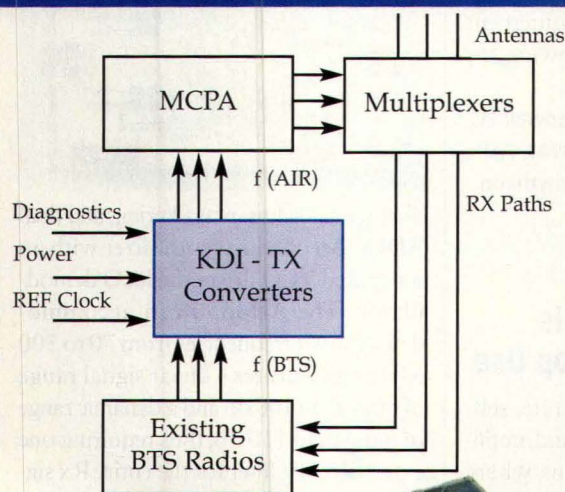
series are designed for operating temperatures from -50 to +100°C and feature insertion-loss repeatability (from one mating cycle to another) of typically ± 0.1 dB. Both connector series are stainless-steel 2.92-mm designs with gold-plated contacts.

The company also offers a lower-cost 2.92-mm stainless-steel version of the front-locking blind-mating connectors in the form of the model 7041, which is rated for 0.6 dB maximum insertion loss and 1.40:1 maximum VSWR from DC to 18 GHz. The blind-mate connectors are ideal for applications requiring the installation and/or exchange of RF modules in mere seconds, without sacrificing electrical performance compared to conventional threaded connectors. The company offers a wide array of options, including interfaces for SMA, SSMA, SMB, 2.4-mm, 2.92-mm, and 3.5-mm connectors. MCE/Weinschel, 5305 Spectrum Dr., Frederick, MD 21703-7362; (301) 846-9222, (800) 638-2048, FAX: (301) 846-9116, e-mail: sales@weinschel.com, Internet: www.weinschel.com.

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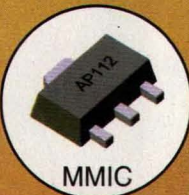
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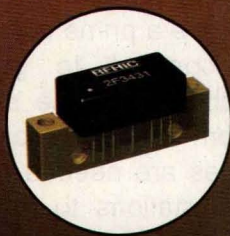


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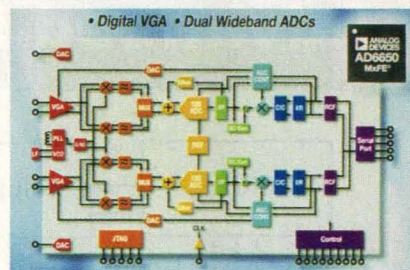
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Receiver Suits GSM/EDGE Wireless Base Stations

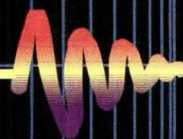
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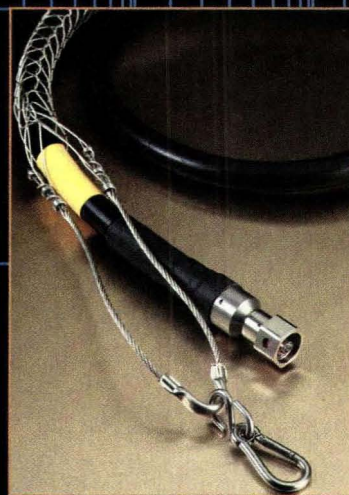
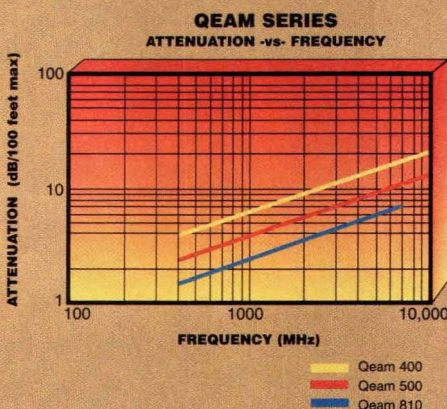
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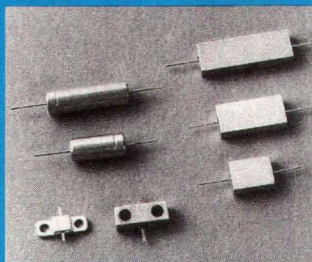
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9dB	DBTC-9-4	5-1000	1.2	18
10dB	DBTC-10-4-75	5-1000	1.4	20
12dB	DBTC-12-4	5-1000	0.7	21
13dB	DBTC-13-4	5-1000	0.7	18
13dB	DBTC-13-5-75	5-1000	1.0	19
		1000-1500	1.4	17
16dB	DBTC-16-5-75	5-1000	1.0	21
		1000-1500	1.3	19
17dB	DBTC-17-5	50-1000	0.9	20
		1000-1500	1.0	20
		1500-2000	1.1	14
18dB	DBTC-18-4-75	5-1000	0.8	21
20dB	DBTC-20-4	20-1000	0.4	21

Dimensions 0.15" square.

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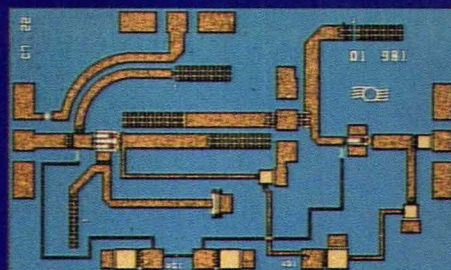
THIS INTEGRATED COAXIAL SWITCH Matrix system with TTL control interface performs up to 18 GHz and is configured as a 2×5 bidirectional, non-blocking Switch Matrix. This unit is suitable for all applications where size and weight are critical. The unit envelope dimensions are $4.3 \times 2.6 \times 1.75$ in. ($10.9 \times 6.6 \times 4.4$ cm), and its maximum weight is 500 g. DC characteristics include 24-to-32-VDC (28 VDC nominal) operating voltage and 760 mA (max) operating current at 28 VDC and 20°C. The 5025 Switch Matrix is designed for broadband signal routing.

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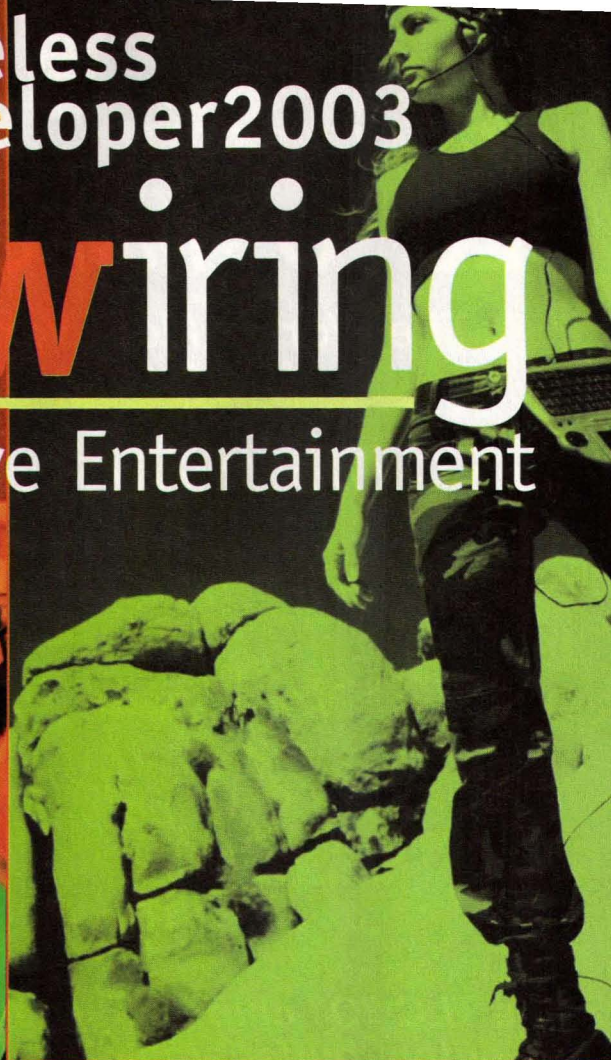
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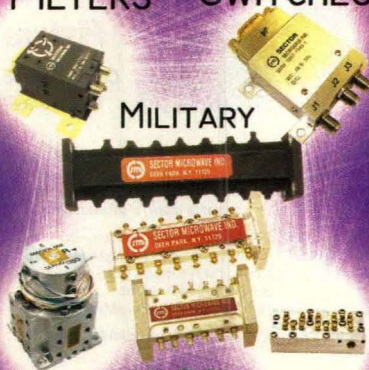


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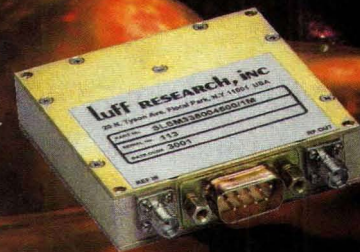


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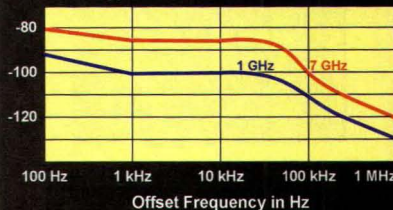
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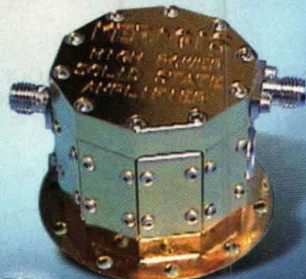
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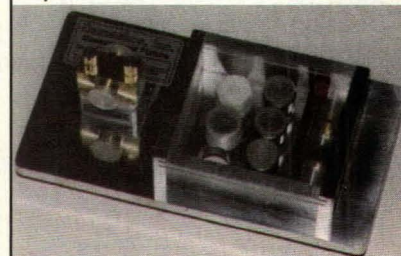


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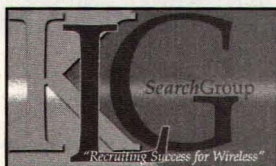


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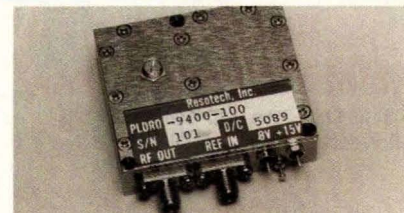
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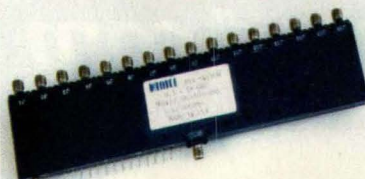
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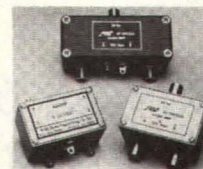
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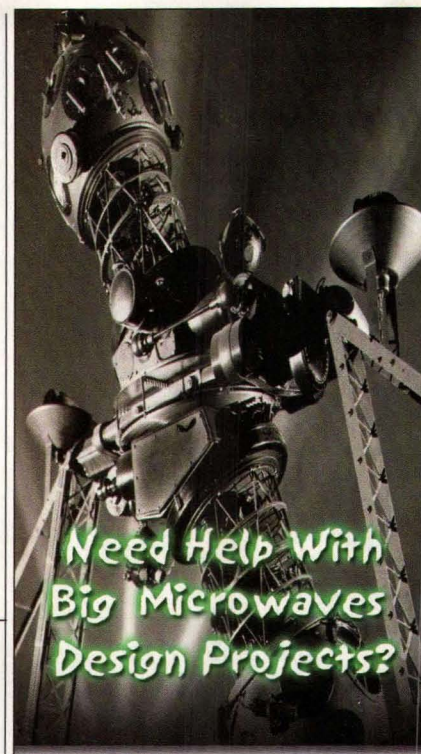
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MARKETING AND ADVERTISING STAFF

GROUP PUBLISHER Craig Roth (201) 393-6225 e-mail: crotth@penton.com SALES ASSISTANT Judy Kollarik (201) 393-6218 e-mail: jkollarik@penton.com DIRECT CONNECTION ADS CLASSIFIED ADVERTISING Joanne Reggas (201) 666-6698 e-mail: jreggas@msn.com CIRCULATION CUSTOMER SERVICE (LIVE) (847) 647-6657 e-mail: microwaves@halldata.com	NEW YORK, NEW ENGLAND, MIDDLE ATLANTIC, DC, VA, SOUTHEAST, MIDWEST, CANADA Paul Barkman Regional Sales Manager Penton Media, Inc. 611 Route #46 West Hasbrouck Heights, NJ 07604 (908) 704-2460 FAX: (908) 704-2486 e-mail: pbarkman@penton.com CALIFORNIA, NORTHWEST, SOUTHWEST Nichole Fox Regional Sales Manager Penton Media, Inc. 45 Eisenhower Dr., fifth floor Paramus, NJ 07652 (760) 476-3447 FAX: (760) 931-9655 e-mail: nfox@penton.com	ITALY Cesare Casiraghi Viale Varase 39 22100 Como - Italy Phone: 39-31-261407 FAX: 39-31-261380 GERMANY, AUSTRIA, SWITZERLAND Friedrich K. Anacker Managing Director InterMedia Partners GmbH (IMP) Deutscher Ring 40 42327 Wuppertal Germany Phone: 011-49-202-271-690 FAX: 011-49-202-271-6920 SPAIN Luis Andrade, Miquel Esteban Espana Publicidad Internacional Sepulveda, 143-38 08011 Barcelona, Spain Phone: 011-34-93-323-3031 FAX: 011-34-93-453-2977	FRANCE Emmanuel Archambeaud Defense & Communication 48 Bd Jean-Jaures, 92110 Clichy France Phone: 33-01-47-30-7180 FAX: 33-01-47-30-0189 HOLLAND, BELGIUM William J.M. Sanders, S.I.P.A.S. Rechtsstraat 58 1483 Be De Ryp, Holland Phone: 31-299-671303 FAX: 31-299-671500 CZECH REPUBLIC Robert Bilek Production International Slezska 61, 13000 Praha 3 Czech Republic Phone: 011-42-2-730-346 FAX: 011-42-2-730-346	PORTUGAL Paulo Andrade Ilimitada-Publicidade Internacional, LDA Av. Eng. Duarte Pacheco Empredimento das Amoreiras-Torre 2 Piso 11-Sala 11 1070 Lisboa, Portugal Phone: 351-1-3863176 FAX: 351-1-3863283 TAIWAN, R.O.C. Charles C.Y. Liu, President Two-Way Communications Co., Ltd. 11F/1, No. 421 Sung Shan Road Taipei 110, Taiwan, R.O.C. Phone: 886-2-727-7799 FAX: 886-2-728-3686 JAPAN Hiro Morita Japan Advertising Communications, Inc. Three Star Building 3-10-3 Kanda Jimbocho Chiyoda-ku, Tokyo 101, Japan Phone: 81-3-3261-4591 FAX: 81-3-3261-6126	KOREA BISCOM Jo Young Sang Rm. 521 Midopa Bldg. 145 Dan Ju-Dong Chongno-gu Seoul 110-071 Korea Phone: 027397840 FAX: 027323662 INDIA Shivaji Bhattacharjee Information & Education Services 1st Floor, 30-B, Ber Sarai Village, Near I.I.T. Hauz Khas, Behind South Indian Temple New Delhi, 110016 India FAX: 001-91-11-6876615
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Current (Amps):	1.0	.90

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Freq. (GHz):	2.0 - 4.0	3.4 - 3.6
Gain (dB):	46.0	48.0
N.F. (dB):	2.0	6.5
Pout (dBm):	20.0	38.0
VSWR (I/O):	2.0:1	1.5:1
Current (Amps):	.260	3.8

MODEL:	MSH-6455402-DI	MSH-5427801
Freq. (GHz):	4.0 - 8.0	6.4 - 7.2
Gain (dB):	26.0	29.0
N.F. (dB):	6.0	8.0
Pout (dBm):	20.0	37.0
VSWR (I/O):	2.0:1	2.0:1
Current (Amps):	.150	3.6

MODEL:	MSH-6544402-DI	MSH-6706805-TC
Freq. (GHz):	8.0 - 12.0	10.15-10.7
Gain (dB):	35.0	48.0
N.F. (dB):	5.0	6.5
Pout (dBm):	20.0	38.0
VSWR (I/O):	2.0:1	1.5:1
Current (Amps):	.250	4.2

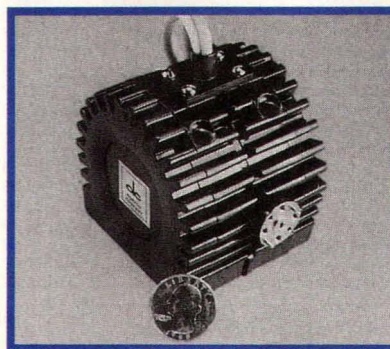
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Freq. (GHz):	12.0-18.0	12.7-13.2
Gain (dB):	21.0	17.0
N.F. (dB):	4.0	2.7
Pout (dBm):	20.0	10.0
VSWR (I/O):	2.0:1	2.0:1
Current (Amps):	.200	.110

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—looking back—



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→next month

Microwaves & RF June Editorial Preview

Issue Theme: Defense/Security Electronics

News

High-frequency companies have renewed interest in defense electronics, due in part to a weakening in commercial wireless and communications markets. Some firms that remained true to these defense markets have maintained fairly steady sales even during recent trying times. What type of technologies are of interest to present-day military customers, and what are the sales expectations for high-frequency suppliers competing in these markets for the next few years? A Special Report in the June issue will attempt to answer these questions and make some accurate projections about the types of hardware, software, and test equipment required by military customers over the next several years. In addition, the June News section will feature a brief wrap-up of some of the key new products from this year's MTT-S exhibition.

Design Features

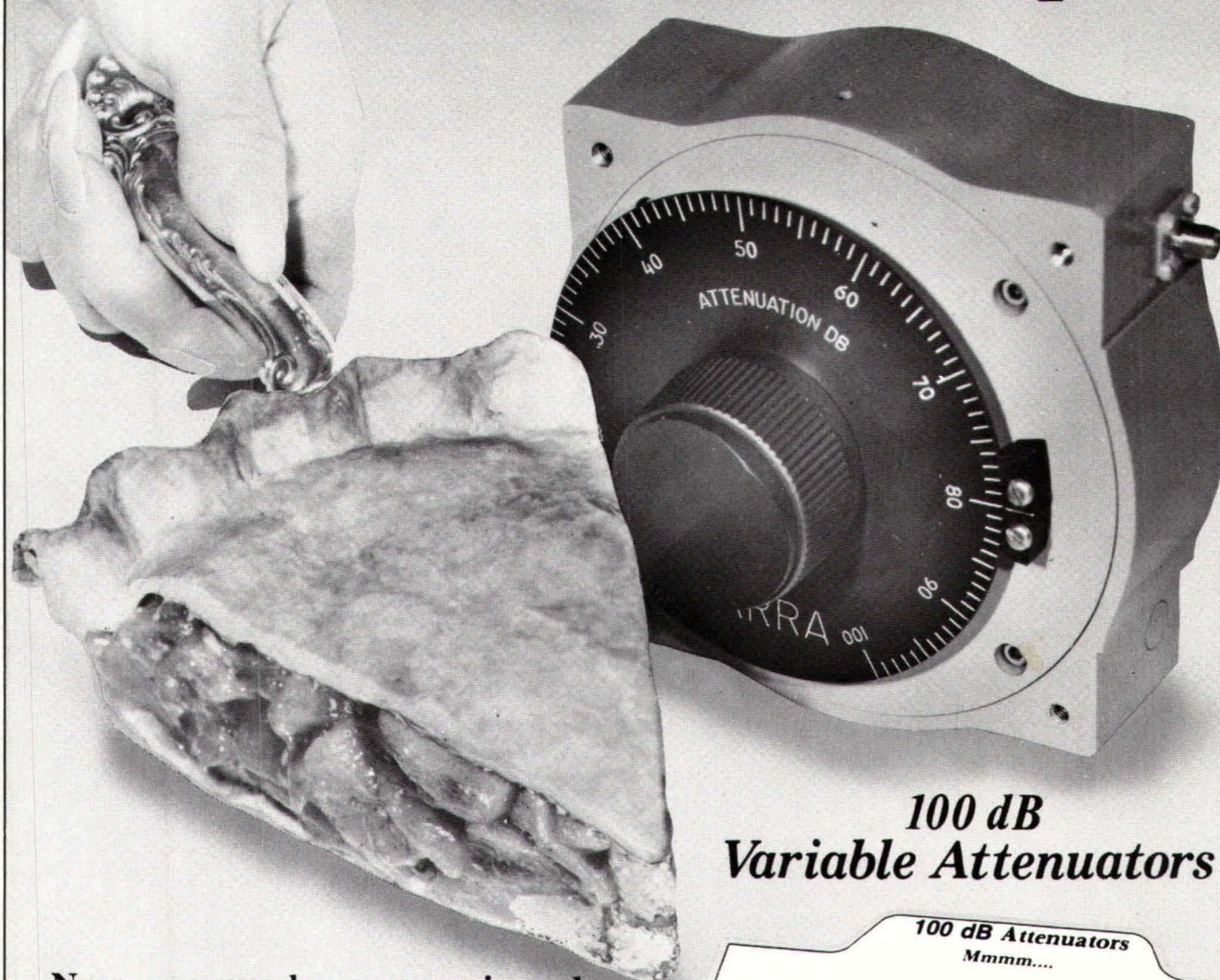
The June Design Features section is brimming with practical technology, including a first look at the Vivaldi antenna, a design aimed at ultrawideband (UWB) communications applications. An author from a leading design software supplier will show how

to use electronic-design-automation (EDA) software to design subharmonic mixers. Also, an author from Paradigm Wireless Systems explores noise-mixing effects in high-power amplifiers (HPAs), while authors from Keithley provide guidelines for constructing an effective RF/microwave switching system. Finally, an author from a top semiconductor supplier will review requirements for third-generation cellular (WCDMA) transmitters.

Product Technology

June offers the first look at a new line of packaged quadrature RF/microwave couplers measuring just 0.15×0.15 in. and offering tight coupling, low loss, and high isolation over broad bandwidths. The Product Technology section also reviews Version 9.0 of a leading three-dimensional electromagnetic (EM) simulation software package, and provides a second look at an innovative analog linearizer that increases power-amplifier efficiency and linearity. Additional stories review a high-performance analog-to-digital converter (ADC) driver, a line of millimeter-wave amplifiers based on novel spatial-combining techniques, and an extremely broadband antenna designed for electromagnetic-compatibility (EMC) testing.

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1000 - 2000 MHz	1.5	3952 - 100X
2000 - 4000 MHz	1.5	4952 - 100 X
4000 - 8000 MHz	1.5	5952 - 100X
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VSWR - 1.5		
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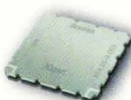
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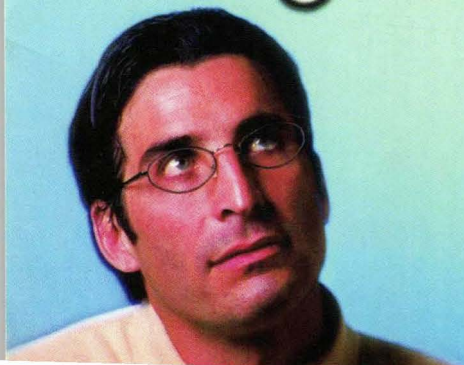


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